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MEETING THE DEMANDS FOR VEHICULAR COMMUNICATIONS*

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The purpose of this paper is to present some of the methods used by the Federal Communications Commission staff in analyzing frequency usage of existing radio services in order to lay the ground work for rule making proposals which would affect the redistribution of frequencies among the various services licensed by the Commission.

Over a period of time, the Federal Communications Commission handles an appreciable number of rule making proceedings dealing with the subject of frequency allocations or arrangements for frequency use. The Commission is constantly striving to obtain the best possible utilization of the existing radio frequencies. However, a large number of the rule making proposals issued by the Commission involves some minor shifting within a basic structure, such as a change in eligibility, or some other minor factor involving the use of frequencies already assigned to a service. On the other hand, over a long period of time there are an appreciable number of rule making proposals made by the Commission that involve fundamental and extensive rearrangements of the use of frequencies. Considerable effort is made to arrive at a rule making proposal, in order to obtain the best utilization of frequencies, and, at the same time achieve a balance among many non-technical factors. In preparation for these proposals, the Commission makes studies and analyses of the use of existing frequencies and potential arrangements to obtain better use of the frequencies. Examples of these larger rule making proposals which involve fundamental changes and which have had considerable study behind them on the part of the Commission's staff before they are proposed formally by the Commission are:

1. The rearrangement of the Standard Broadcast frequencies between 1938 and 1941 that led up to the present North American Regional Broadcasting Agreement, which was adopted in 1941.

2. The Television Allocation Plans from 1946 to 1950.

3. The Commission's proposals to utilize additional channels created by splitting the existing frequencies for land mobile use.

4. All of the Commission's overall basic allocation hearings where major redistributions of frequencies among the existing users was proposed to prepare the United States position for the Radio Conferences of the International Telecommunications Union.

One basic tool used by the Commission in analyzing the use of radio frequencies is the channel study map. A mark is made on a map for each existing or proposed assignment. This mark may or may not have direct correlation with the technical operating characteristics of the station. For example, an "X" on the map at a certain place will indicate that a station is located at that place and no more, whereas, in producing the television allocation plan, a circle is used which would be approximately one-half of the normal co-channel separation for two stations of equal power and equal antenna height, the power used being the maximum allowed by the rules, the antenna height used being a value slightly above average. Somewhat similar techniques are used in the land mobile service studies by using a circle that is approximately the service area radius.

In some of the broadcast services, these maps are kept up on a continuous basis, due to the mutually exclusive nature of the applications filed. Whereas, in most other services they are only made up on a spot basis to analyze an existing situation that is involved in channel occupancy. At a glance, such maps show a number of things. For instance, (1) the efficiency of utilization of a particular channel or (2) where additional assignments may be easily made without major conflict with other assignments. They may also show that a certain industry's use of frequencies does not necessarily peak at the same place that another industry's use of frequencies may peak. In using these analyses that may be derived from the channel study map, it is necessary to correlate the circles on the map with something that may be used as a projection tool. It is these correlations that point the way to certain proposals. It would be an ideal situation if the use of radio frequencies were directly related to area. Of course, we all know that that is not true, as the use of frequencies tends to be heaviest in the large cities like New York, Chicago and Los Angeles. It is more logical to consider that the use of radio frequencies are proportional to population distribution. However, frequency channel studies show this is not true either. The next matter for consideration is to discover which industries' use of radio frequencies do follow population distribution and which do not. If the use does not follow population distribution, it is necessary to determine the characteristic involved that may be useful in making rules. The studies over a period of time have shown us a number of useful facts. The following industries'use of radio frequencies do not follow population distribution:

*Presented at the PGVC National Conference in Washington, D. C., December, 1957.

Marine use of radio frequencies follows our coastline primarily with some secondary use along the rivers, but the use along the rivers is not as great as that along the coastline.

The Railroads' use of frequencies tends to peak in the Chicago area, rather than in the New York or Los Angeles areas, and the use of frequencies is not proportional to the population of any particular city, but is nearer in proportion to the number of railroads that may be going in and out of any particular city.

Forestry Conservation's use of frequencies is, of course, largely associated with the particular percentage of a state that uses forestry --that is forested. Although all States have forestry conservation radio, there is a general tendency for the use and the number of frequencies desired to be much heavier in the States with very high percentages of forest areas and where the forest industry also operates in the same State rather extensively. For example, this is true in Washington, Oregon and Maine.

The Forest Industry's use of radio, which is the commercial counterpart of forestry conservation, very definitely follows the forestry industry. Where the forestry industry is a major industry in Washington, Oregon and northern California, the forest industry's use of radio peaks in that area.

The Petroleum Industry's use of radio frequencies and its various satellite industries serving the industry peak where oil production is the greatest along the Gulf Coast from Florida to the Mexican Border, through Louisiana and Texas, in west Texas and the eastern edge of New Mexico, with a lesser use of frequencies in other parts of the country.

The Mining Industry's use of radio might appear to be in almost all States of the union. However, certain aspects of the mining industry, such as coal mining, have not found radio an effective tool. Therefore, there is a tendency for radio to be used by the mining industry throughout the mountain States, from Arizona, New Mexico, Nevada, Utah, Colorado, Wyoming, Idaho, and Montana, and the iron mining industry in the Minnesota area, with little use throughout the industrial east of the United States or near the West Coast.

Agricultural use of radio is almost directly opposite to population distribution in that it tends to follow the large commercial farming areas of the country, especially, in the Sacramento Valley of California and in the State of Iowa.

Taxis' use of radio frequencies might be said to follow population distribution accentuated. The taxis' use of radio tends to be in the large metropolitan areas and their use drops off very rapidly in proportion in towns and cities below 50,000 population. It is definitely a "big-town" use of frequencies.

The Transit Industry's use of radio is probably the most accentuated in that aspect. A recent analysis shows that, with only two exceptions, the transit industry's use of radio is confined to the large metropolitan areas using public transit systems of considerable size.

When these analyses of each industry's use of radio versus population distribution have been made, it is then possible to set up combinations that will give the maximum frequency utilization. By associating the peaks of use that do not follow population distribution with those that do follow population distribution, a "pool" of frequencies is established in which the maximum utilization is achieved with the minimum of conflict between users. Commission Rules have had the forest industries and the petroleum industry sharing the same frequencies for a number of years; however, the use of the frequencies has not been as effective as hoped for as neither group uses the shared frequencies effectively in the industrial eastern part of the United States. Consequently, the Commission proposed the manufacturers as a third sharing industry on the frequencies already assigned to the petroleum and forest products industries.

One might carry the reasoning further and create a large "pool" of all the frequencies assigned to land mobile use. The idea has many attractions, especially to users in a single industry area or in sparsely populated areas. The Commission's Rules have overcompartmented the use of frequencies with the result that several industries in sparsely settled areas are confined to less than 10% of the existing frequencies in areas where there is little or no use for the other 90%. It is in situations such as this that even small pools of frequencies become of immense value as the total frequency availability is doubled or tripled for most industries. There are, however, disadvantages to a large pool of frequencies which is operated on a first come, first served basis. Some users of radio frequencies, such as railroads, petroleum pipelines, and state police have a ribbon system or wide area problem. Where the use of frequencies is to be over a wide area and the system is to grow by expanding the area of operation, it is both convenient and practical to either plan for the use of one frequency for the wide area system, or allow enough flexibility in the number of frequencies assigned to a given industry to accommodate the wide area problems with the minimum of impact on existing systems. The advantages of advance planning to meet this situation have been so apparent for a number of years that all frequencies assigned for use by the States are preplanned for the most effective use, while still allowing for future growth. If all land mobile frequencies were in a large pool on a first come, first served basis, wide area users would have to use multichannel equipment to accommodate the different frequencies from area to area, or to regard each as a separate area problem and establish a universal calling frequency for all areas, and individual working frequencies for each area.

In analyzing the various uses of radio frequencies after they have been in existence, the growth rates must be determined. If new frequencies are to be derived or frequencies taken away from some service, it is necessary to know, not only the density of use per channel, but also whether the particular use is in the explosive stage of growth, or is in the later stage of growth. The Commission has plotted the growth rates of the various industries' use of radio. In the industrial field and in the motor trucking field the growth rates are still in their explosive stage, whereas, in taxicabs, fire and police, the growth rates have slowed down. Therefore, in making any proposal where there are rearrangement of channels, these growth rates have to be taken into consideration. This was done when the Commission put out its split channel proposels. In the Commission growth rate studies it was decided from the growth trends that certain industries needed an expansion of two or three times the number of frequencies assigned to them, whereas, other industries merited little or no expansion on the number of frequencies already assigned them.

Overall, the use of land mobile radio frequencies is still growing rapidly with some industries being in the explosive stage of growth, whereas, a few industries have probably reached the maximum stage of growth. It appears that land mobile use of radio will continue to grow for an appreciable period of time. The Commission has proposed channel splitting to increase the number of available frequencies, and at the same time, has achieved better geographical sharing by the creation of frequency pools available to all users and by regrouping the industries whose use of radio frequencies is peaked in one region of the United States. The rearrangements proposed, if implemented, should allow for the continued expansion of land mobile radio use for a period of at least five years.

INTERFERENCE -- A LOOK AT THE OUNCE OF PREVENTION*

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Summary

One of the major interference problems in the field of vehicular communications is brought about by the concentration of a number of base station installations in a relatively small area, or even at the same antenna site. Most cases of interference resulting from such multiple installations are difficult to analyze after the fact. A method is described, herein, for predicting such interference before it occurs. Futhermore, as a method for alleviating radio interference problems due to frequency and geographic congestion, a technique is suggested for processing new applications for frequency assignment.

There have been so many papers written and presented on the subject of interference due to spectrum crowding, that it has become trite to state that this is the most serious problem facing vehicular communications today. Nevertheless, the problem becomes more accute with each passing day. Every new installation is regarded with apprehension by existing licensees as another possible source of interference. In a great many cases, equipment is installed only after considerable thought and calculation has gone into the selection of a frequency and a location. Unfortunately, there are also quite a few licensees that romp merrily into a new installation, blissfully unaware of the complexity of the situation. All too frequently, these latter cases result in the laborious task of tracking down a source of interference either to the new licensee or one of his radio neighbors.

Determining the source of interference, after the fact, is a time consuming and complex task. Only the most careful evaluation of the interference can determine whether a receiver or a transmitter is at fault. The fault is difficult to establish, since similar interference symptoms will occur due to the limiting performance of the receiver and extra band radiation of an interfering transmitter.† A somewhat more straight-forward job is that of predicting interference before it occurs. This is the ounce of prevention. However, I do not mean to imply that the job is a simple one.

When analyzing a proposed new installation, or a proposed new system, possible interference must be considered in two general categories (Chart 1). In the first category are those causes of interference caused by the limiting performance of the new receiver, or of existing receivers at the same site. The first case is intermodulation produced in the front end of the receiver, due to two or more transmitters operating nearby on closely spaced frequencies. The second is modulation splatter from an adjacent channel transmitter that might be heard in the newly installed receiver, due to the limitations of the receiver selectivity. A third possibility is desensitization of the desired signal due to a strong nearby undesired signal, and a fourth is interference due to spurious and image responses of the newly installed receiver.

CATEGORY I

Interference caused by limiting performance of the receiver.

1. Intermodulation
2. Modulation Splatter
3. Signal Desensitation
4. Spurious and Image Response

CATEGORY II

Interference caused by extra band radiation of the interferring transmitter.

1. Intermodulation
2. Modulation Splatter
3. Broadband Noise Radiation (Transmitter Noise)
4. Spurious and Harmonic Radiation

The second category covers possible interference caused by extra band radiation of an interfering transmitter. In this category we must also consider interference that a newly installed transmitter might cause to existing installations at the same site. The first such source of interference in this category is intermodulation produced in the final stages of simultaneously operating transmitters operating in close proximity to each other. The second source, as in category one, is modulation splatter. However, in this case, it is due to the limiting performance of the transmitter and may be due to over-modulation. The third source is broad band noise radiation - essentially, a normal characteristic of a power amplifier. This is usually termed "transmitter noise". Finally, we must consider spurious and harmonic radiations from the transmitter.

To predict the level of each type of interference for all categories, one requires the complete interference characteristics of all

*Presented at a meeting of the Washington Chapter of the PGVC in May, 1959.

†Shepherd, N. H.: "Bi-dimensional interference analysis." Proc. Third Conf. on Radio Interference Reduction.

transmitters and receivers in a given situation, plus the attenuation values between antennas at the proposed installation, plus the operating frequencies involved. Only when all of these factors are known, can a really accurate prediction be made. Of course, in a great many cases, all variables are not known and some guesswork is necessary. However, even this is preferable to walking blindly into a new installation with no thought to possible interference.

Once he has analyzed a given situation, the application engineer may utilize three basic dimensions in order to arrive at a condition of minimum interference. These dimensions are time, space, and frequency. The effects of the dimension, time, are probably not quite as obvious in the prediction of interference as are the effects of the other two. However, in extremely crowded areas, it may be equally as important. For instance, let us suppose that there are three transmitters within interfering distance of the proposed new site. The combination of these three transmitting frequencies may produce an inter-modulation product which will interfer with the proposed newly installed receiver. However, if the duty cycle on each of these three transmitters is light, the probability of all three being on simultaneously is not very great, and, therefore, there will be little interference to the proposed installation. On the other hand, if one or more of these transmitters operates with a continuous carrier, interference will occur far more frequently, and this may be predicted with a fair degree of accuracy. In other words, it is not enough to know what interference may occur. The seriousness of the interferences must also be determined, and then a decision made as to whether it can be tolerated for a given situation.

Now let's consider the dimension, frequency. It would be a fortunate applicant indeed who had his choice of more than two or three frequency assignments out of the total mobile spectrum. Nevertheless, it is to everyone's best interest, including his own, that he make the best choice. This brings us to what is probably the most important and, at the same time, the most tedious part of the process of interference prediction. A thorough analysis should be made of the existing equipments in operation in the general area of the proposed new installation. As an illustration, we will take a hypothetical case.

Let us assume that the problem is to select a set of high band frequencies, one transmitting and one receiving. The frequencies are to be used by a base station, to be installed in a geographical area which has already been heavily congested with equipment operating in the land mobile service. Exhibit AA-1 shows the geographical arrangement of the presently installed equipment. A, B, and C are the sites of presently installed equipment operating in the land mobile service; D is the site of a 5 kilowatt broadcast station operating on 880 KC; E is the site of a 5 kilowatt FM broadcast station operating on 102.2 MC; and F is the site of a 50 kilowatt T.V. station operating on channel 7. The new base station is to operate two-frequency simplex, with a transmitter power output of 40 watts. The transmitter frequencies available for the new base station are 153.47 MC, 153.65 MC or 153.71 MC. The receiver frequencies available

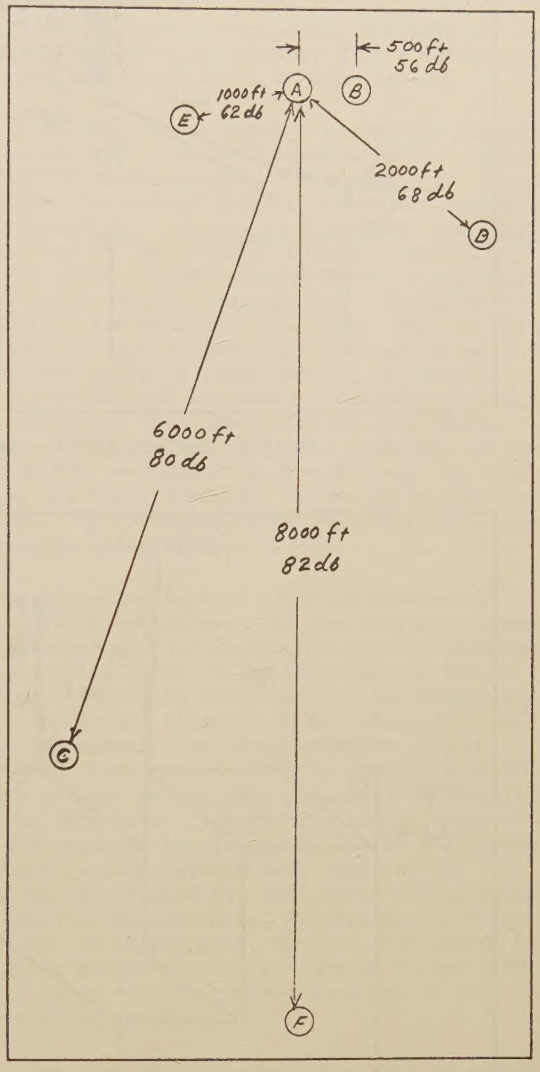
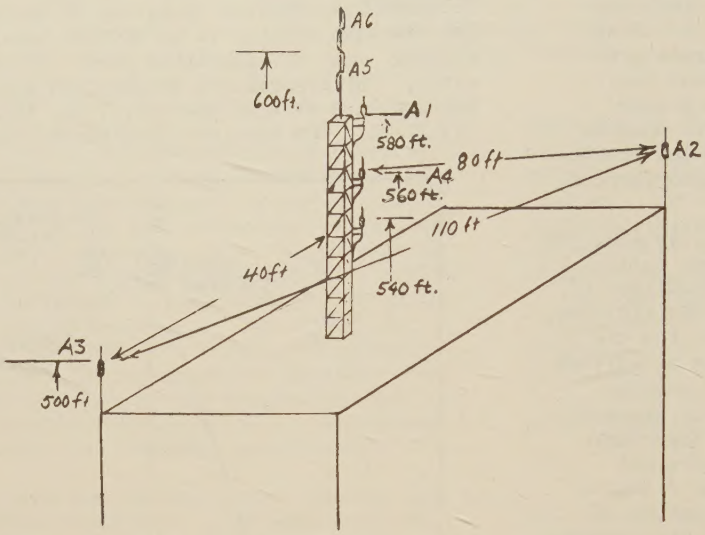


Fig. 1 (Exhibit AA-1) - Station locations.

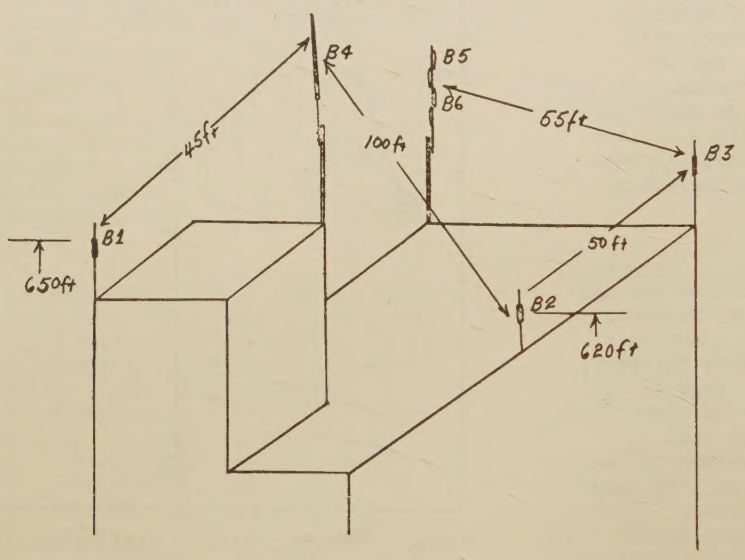
are 158.13 MC or 158.49 MC. It is desired that no greater than 1 db loss of received signal be caused by interfering transmitters when the new receiver is installed at location A. We will assume that there is an average ambient noise level of 1 microvolt at the receiver input. Interference to existing equipment should not create greater than 1 db loss in sensitivity, based on the same assumptions. Exhibit AA-2 shows the antenna arrangements and operating frequencies of the existing equipment at location A. Exhibit AA-3 shows the arrangement at location B and Exhibit AA-4 shows the arrangement at location C.



LOCATION	TRANSMITTER FREQ
A1	152.03MC
A2	157.47MC
A3	161.49MC
A4	152.27MC
A5	457.55MC
A6	460.65MC

RECEIVER FREQ	POWER OUTPUT
158.49MC	40 Watt
157.47MC	30 Watt
161.49MC	30 Watt
157.53MC	40 Watt
452.55MC	20 Watt
460.65MC	20 Watt

Fig. 2 (Exhibit AA-2) - Antenna arrangement at base station A.



LOCATION	TRANSMITTER FREQ
B1	156.03MC
B2	154.19MC
B3	153.17MC
B4	152.39MC
B5	460.85MC
B6	453.25MC

RECEIVER FREQ	POWER OUTPUT
156.03MC	200 Watt
154.19MC	200 Watt
153.17MC	200 Watt
157.65MC	40 Watt
460.85MC	20 Watt
453.25MC	20 Watt

Fig. 3 (Exhibit AA-3) - Antenna arrangement at base station B.

It may appear that we have made the problem overly complicated. However, this situation is typical, and a similar problem could be encountered in any large city in the country.

The first step is to compute the primary intermodulation frequencies that will be generated at the 3 antenna sites, and compare them with the proposed new frequencies to determine whether there is a possible interference.

Third order IM should be examined for all transmitted signals which are within one mile of each other. In addition, a look should be taken at 5th and 7th orders for all transmitters which are spaced less than 500 feet apart. In the example we have chosen, we have IM products coming out of our ears. To illustrate, examples of third order IM are calculated below for each station location and examples of station to station IM are given. One examples of 5th order

IM is included. The IM frequencies generated at station A are circled below:

STATION A IN MC

Tran.	Frequency MC		3rd Order	5th Order
A1	152.03		(151.79)	(151.55)
A4	152.27	.24	(152.51)	(152.75)
A1	152.03	5.44	(146.59)	
A2	157.47		(162.71)	
A1	152.03	9.46	(142.57)	
A3	161.49		(170.95)	
A4	152.27	5.20	(147.07)	
A2	157.47		(162.67)	
A4	152.27	9.22	(143.00)	
A3	161.49		(170.71)	
A2	157.47	4.02	(153.45)	
A3	161.49		(165.51)	
A1	152.03		(146.98)	
A4	152.27		(157.23)	
A2	157.57		(157.71)	
A4	152.27		(148.25)	
A2	157.47		(156.29)	
A3	161.49		(166.69)	
A2	157.47		(156.05)	
A3	161.49		(148.01)	
A1	152.03		(166.93)	
A1	152.03		(161.73)	
A4	152.27		(161.25)	
A3	161.49		(142.81)	

STATION B & C

In a similar manner the third order IM products from Stations B & C can be calculated and are listed below:

148.05	152.03		157.47
148.75	152.27	153.95	157.47
149.53	152.35	154.07	157.87
150.31	152.39	154.59	158.89
150.33	152.51	154.93	158.91
150.58		155.01	159.67
	152.75	155.21	160.01
151.33	152.87	155.25	
151.37	152.99	155.99	
151.61	153.17	156.03	
151.67	153.41	156.81	
152.15		157.05	

Out of all of the third order products generated at the three station sites, there is one possible interference. One of the undesired frequencies generated by the combination of transmitter A2 and A3 lands on 153.45 MC, which is 20KC removed from the proposed transmitter frequency 153.47 MC. However, the two fundamental frequencies are separated by 4.02 MC. Unless a mobile unit was within 500 feet of station A, it is unlikely that any interference would be caused. Those of you who examine the list of intermodulation frequencies will note that several of them fall on the assigned frequency of some of the existing equipment. One example is the frequency A2 (157.47 MC), created by the combination of transmitters C1 and C2. This type of interference could probably be reduced by inserting a cavity filter between the antenna and the RF input of receiver A2.

We should also examine second and fourth intermodulation products created by the combination of nearby broadcast transmitters. In our example, the second order products are sufficiently removed from the proposed receiver frequencies so as not to cause interference. However, a fourth order product created by twice the difference of the FM broadcast station frequency and the T. V. frequency falls at 155.1 MC. At first glance, it would appear that this could be neglected since it falls about 3 MC from our proposed receiver frequency. However, it should be remembered that the bandwidth of a fourth order product is two times the bandwidth of the sum of the two producing frequencies. In this case, the total bandwidth of the IM product is 9.2 MC. The resulting broadband noise spectrum may cause interference.

After the various existing sources of possible IM are checked out and certain proposed frequencies are eliminated, it is then necessary to determine which of the proposed new transmitter frequencies may create IM interference, if added to the system. By so doing, we find that the combination of transmitter B3 and proposed frequency 153.71 will produce a third order product on the same frequency as receiver C1. By the same token, the combination of transmitter B3 and proposed new frequency 153.65 will produce a possible interference near receiver B2. We have now eliminated all but one of the proposed transmitter frequencies. Futhermore, by examining the list of IM products, we find that one of the receiver frequencies is preferable to the other since it is further removed from one of the possible IM products.

Now that we have selected a pair of operating frequencies, the next step is to determine the proper physical spacing between antennas at site A. Exhibit A5 illustrates the type of curves necessary to make this determination. Essentially, what these curves show is the necessary attenuation between a transmitter of known characteristics as we vary the spacing between the transmitter frequency and the receiver frequency. In order to apply these curves to the problem at hand, we need to know the minimum frequency spacing between any of the presently installed transmitters at site A and our proposed receiver.

Proposed receiver frequency	158.13 MC
Transmitter frequency A2	157.47 MC
	f .66 MC

At the same time, we need to know the frequency spacing to the nearest frequency at site A from the proposed transmitter. The latter is done so that we can claim that we live by the golden rule as well as to avoid distasteful occurrences, like being lynched at midnight by the other licensees using site A.

Receiver frequency A2	157.47 MC
Proposed transmitter frequency	153.65 MC
	f 3.82 MC

The curves show that, for a frequency separation of .66 megacycles, we require 55.6 db of attenuation between antennas, in order to loose no more than 1 db of the receiver sensitivity due to the effects of transmitter noise. This brings us to a rather sticky problem. In order to obtain this much attenuation, we need to space our new receiving antenna at least 30 feet vertically or 230 feet horizontally from the antenna for transmitter A2. There just isn't that much elbow room at location A without new construction and rather expensive relocation of existing antennas. The only other alternative is to provide a cavity filter for transmitter A2.

This is all very nice, but who is going to finance the purchase and installation of the necessary cavity filter - the new licensee, or the owner of transmitter A2? Let's assume that one or both of the parties involved is more far-sighted than most of us, and the cavity is installed. With a $\frac{1}{2}$ db coupling loop installed, the cavity will provide 16 db of attenuation at 660 KC away from its center frequency. Now we need only about 40 db between antennas which can be obtained with a horizontal spacing of 36 feet.

Next, we take another look at the curves to see what attenuation is going to be required between our proposed transmitter and receiver A2. We see that this required attenuation is 32 db, and that it can be obtained with a vertical spacing of 8 feet or a horizontal spacing of 15 feet. Since we have already decided to space our new antenna a horizontal distance of 36 feet from antenna A2, this is no problem.

As mentioned earlier, another type of interference that must be considered, in an installation of this type, is the spurious output of the

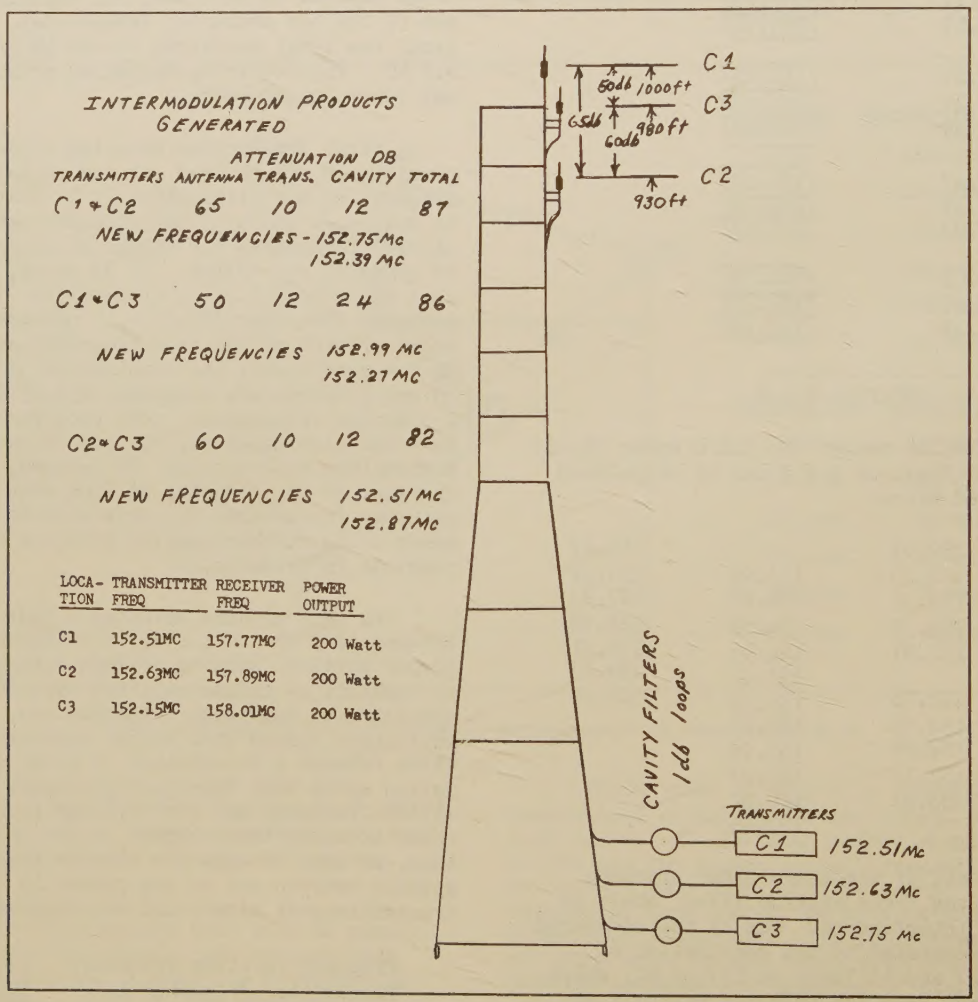
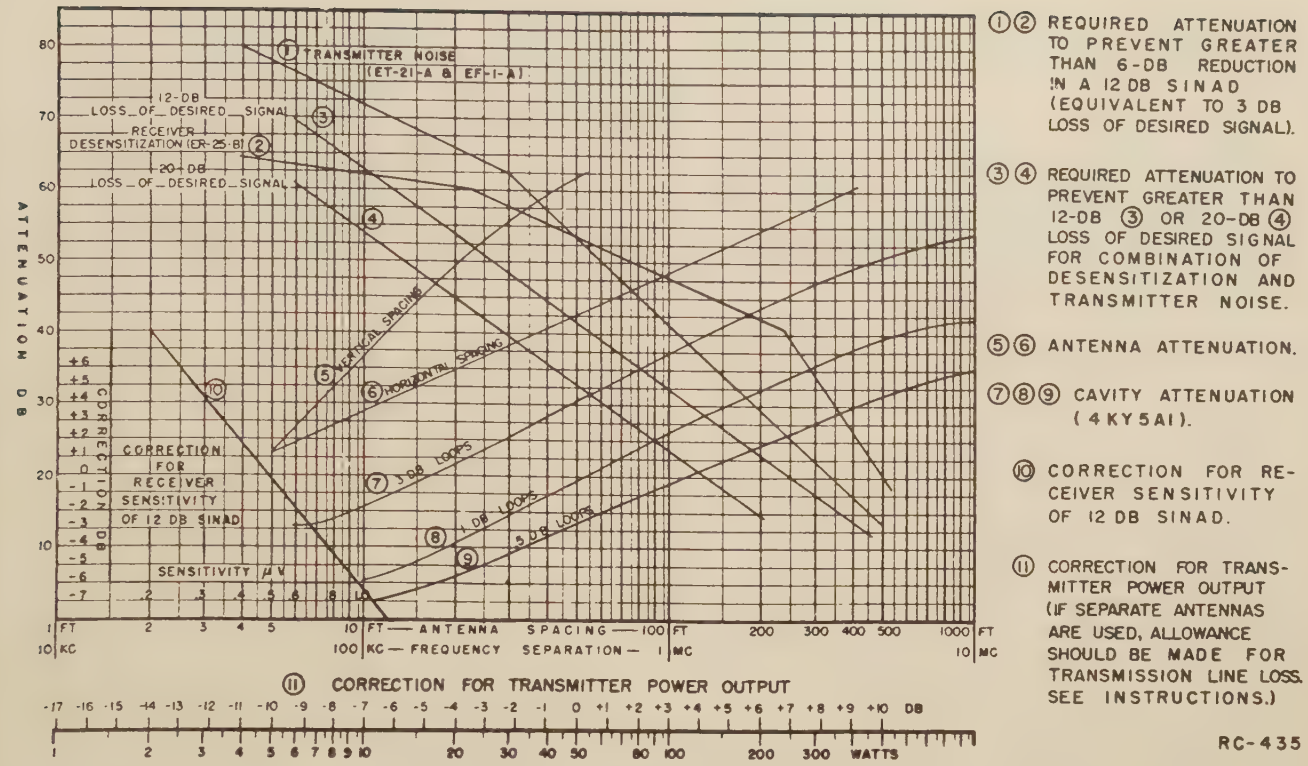


Fig. 4 (Exhibit AA-4) - Antenna arrangement at base station C.


① TRANSMITTER NOISE (ET-21-A & EF-1-A)
12-DB LOSS OF DESIRED SIGNAL
RECEIVER DESENSITIZATION (ER-25-B)
②
20-DB LOSS OF DESIRED SIGNAL
③
④
⑤ VERTICAL SPACING
⑥ HORIZONTAL SPACING
⑦ 3 DB LOOPS
⑧ 1 DB LOOPS
⑨ .5 DB LOOPS
⑩ CORRECTION FOR RECEIVER SENSITIVITY OF 12 DB SINAD
SENSITIVITY μV
CORRECTION DB
ATTENUATION DB
1 FT — 10 FT — ANTENNA SPACING — 100 FT — 1000 FT
10 KC — 100 KC — FREQUENCY SEPARATION — 1 MC — 10 MC
⑪ CORRECTION FOR TRANSMITTER POWER OUTPUT
-17 -16 -15 -14 -13 -12 -11 -10 -9 -8 -7 -6 -5 -4 -3 -2 -1 0 +1 +2 +3 +4 +5 +6 +7 +8 +9 +10 DB
1 2 3 4 5 6 7 8 9 10 20 30 40 50 80 100 200 300 WATTS

①② REQUIRED ATTENUATION TO PREVENT GREATER THAN 6-DB REDUCTION IN A 12 DB SINAD (EQUIVALENT TO 3 DB LOSS OF DESIRED SIGNAL).

③④ REQUIRED ATTENUATION TO PREVENT GREATER THAN 12-DB ③ OR 20-DB ④ LOSS OF DESIRED SIGNAL FOR COMBINATION OF DESENSITIZATION AND TRANSMITTER NOISE.

⑤⑥ ANTENNA ATTENUATION.

⑦⑧⑨ CAVITY ATTENUATION (4 KY 5 AI).

⑩ CORRECTION FOR RECEIVER SENSITIVITY OF 12 DB SINAD.

⑪ CORRECTION FOR TRANSMITTER POWER OUTPUT (IF SEPARATE ANTENNAS ARE USED, ALLOWANCE SHOULD BE MADE FOR TRANSMISSION LINE LOSS. SEE INSTRUCTIONS.)

RC-435

Fig. 5 (Exhibit A-5) - Typical duplex operation curves (160 mc).

transmitters in the area. One example would be that of 450 megacycle transmitters which employ a tripler as a last multiplier. These transmitters create a spurious output which is one third the output frequency and lands within the 150 MC band. Even though this spurious is well within requirements of the FCC rules, it can cause serious interference to high band receivers located within a radius of 1 mile. In our example, we should take each of the assigned 450 MC frequencies at the various locations and divide the assigned frequency by 3. We then should take a close look at the resulting high band frequencies and be certain that none of them fall sufficiently close to one of the receiver frequencies at the three sites to cause interference. This leaves only the problems of receiver spurious and image response, and modulation splatter to be considered. However, I believe we have sufficiently established a method and an approach for interference prediction, and there is no need to go into the problem further at this time.

In a situation as complex as the one outlined in our example, it is altogether possible that, even after a thorough analysis of this situation, the equipment may be installed and still encounter some interference. For example, in our hypothetical case, we overlooked one possible IM product - that of the combination of transmitters A1 and A4 when combined with the signal of the AM station. These three signals, combined, create an interference on the receiver frequency, 158.19 MC. However, a simple trap, tuned to the AM broadcast frequency, could be installed to reduce interference to an acceptable level.

As you can see from our example, the engineering of new frequencies into a geographical area, already heavily congested with radio communication equipment, is a complex and time-consuming job. And the complexity of the system engineering increases very rapidly as each new frequency is added. On the other hand, such an analysis must be made by someone, if new equipment is to be added in congested areas without creating chaos.

Up until the present, the only way to service the increasing demand for new frequency assignments, has been by the addition of manpower. It would appear that the techniques and the hardware are now available to handle this increasing load without additional manpower -- that is, by the use of a computer. Data concerning the interference characteristics of known receivers and transmitters, the characteristics of antennas and filter devices, and space attenuation in the applicable parts of the frequency spectrum, could be programed on the computer in advance.

When application for a new frequency assignment was received by the Federal Communications Commission, additional information would be fed to the computer, such as, the distribution of noise level and presently assigned frequencies at the proposed new site, plus the available frequencies for assignment. The system would

then select the best possible frequency assignment, as well as furnish a prediction of the level of interference to be expected at the new installation.

Most of the required data, concerning the characteristics of communications equipment and devices is already available. The computer know-how for such a system also exists. The only remaining ingredient is the initiative to establish means of evaluating new frequency assignments in this manner, on a centralized, coordinated basis. The application of data processing techniques to the problem of frequency assignment would permit rapid, accurate and equitable frequency assignment, with concomitant reduction in interference problems and savings in Engineering man-hours. The end result would benefit the Government, industry and the general public in better utilization of the Radio Frequency Spectrum without making unreasonable demands on technical manpower.

OPERATION OF CLOSE-SPACED ANTENNAS IN RADIO RELAY SYSTEMS*

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Summary

The required attenuation between transmitter output and receiver input for any given pair of equipments may be measured by a method that takes into account both receiver desensitization caused by front end overloading and the interference effects of transmitter noise side bands. These are the two principal causes of degradation in receiver performance where simultaneous transmission and reception is to be accomplished from antennas to be placed in close proximity. With the required isolation known, curves may be consulted to determine the necessary antenna physical spacing, either colinear or in the same plane. The actual attenuation between two antennas including their associated transmission lines may be measured and, lastly, the presence of any receiver desensitization may be determined under practical operating conditions. A method for increasing the isolation between close-spaced vertical antennas for a given transmitter operating frequency is described.

Simultaneous Transmission and Reception

Use of radio relays or automatic repeaters in a system requires that simultaneous transmission and reception be satisfactorily accomplished in order that a received signal may be rebroadcast. This introduces a number of problems where for economic, or other reasons, the transmitter and receiver are to be located at the same point. The receiver will be subject to the interference effects of the transmitter with possible degradation in performance commonly called desensitization. The interference effects may be from receiver spurious responses, transmitter spurious emissions, intermodulation products, transmitter noise side bands, and receiver front end overloading. These are by no means the only sources of degradation in receiver performance but cover the majority of which front end overloading and transmitter noise side bands are the principal offenders. At frequency spacings of less than about 200 to 500 kilocycles, transmitter noise side bands predominate; at greater spacings the effect is primarily one of loss of gain in the receiver due to overloading. Transmitter noise side bands have an amplitude that roughly follows the selectivity curve of the final amplifier plate tuned circuit with some variation[1]. The result is the emission of energy within the receiver acceptance band making additional receiver selectivity ineffective. In such cases, a cavity in the transmitter output may be beneficial. At greater frequency spacings, a cavity will be of greatest benefit in the receiver input. In cases where the effects of transmitter noise side bands and receiver overloading are about equal, eliminating or reducing one without materially altering the other will result in no substantial benefit. In any event, there is a minimum attenuation between transmitter output and receiver input below which degradation of receiver performance will be observed.

Measuring Required Isolation

The required attenuation between a transmitter output and receiver input to permit satisfactory simultaneous operation may be measured by a method similar to that described for transmitter noise side band level in EIA Standards RS-152. The method of measurement consists of terminating the radio relay receiver in a standard output load and a distortion analyzer incorporating a 1000 cps filter. A signal generator is coupled to the receiver through an isolating pad in parallel with an adjustable attenuator connected to the transmitter output. The signal generator is adjusted to the receiver operating frequency and modulated at two-thirds of the rated system deviation at 1000 cps. The output is set to a level to produce the standard 12 db signal + noise + distortion to noise + distortion ratio at the receiver output. The transmitter, modulated with two-thirds of maximum rated system deviation at 400 cps, is then turned on and the attenuator adjusted until the 12 db SINAD ratio is decreased to 6 db. The required attenuation will be the attenuator indication plus any insertion loss. The automatic repeater audio connection between the receiver and transmitter must be disconnected or disabled during the test. A block diagram of the necessary connections is shown in Figure 1.

Field measurements may be performed requiring less test equipment if 20 db

*Presented at a meeting of the Washington Chapter of the PGVC in November, 1958.

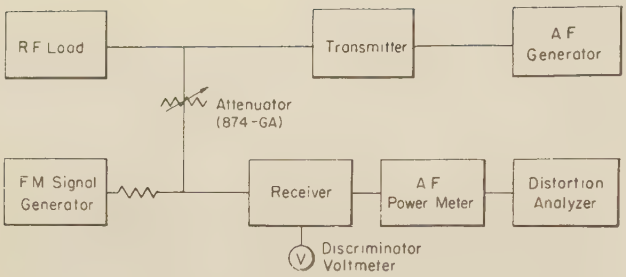

Fig. 1 - Connections for measuring required attenuation between transmitter and receiver using 12 db SINAD reference.

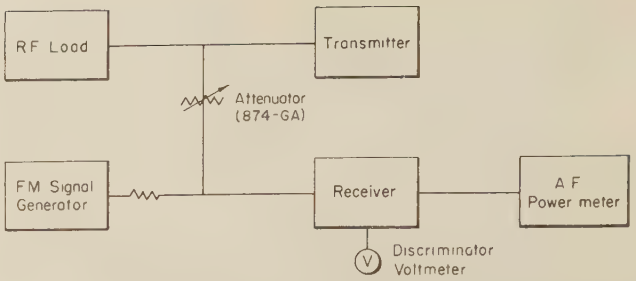

Fig. 2 - Connections for measuring required attenuation using 20 db noise quieting reference.

noise quieting is used as a reference level. The overall results, while not considered as accurate, do not differ greatly from those using the 12 db SINAD as a reference. The signal generator is adjusted until 20 db noise quieting is obtained at the receiver output with the transmitter off. The transmitter is then turned on and the attenuator adjusted until there is a decrease in the noise quieting to 10 db. As before, the required attenuation is the attenuator reading plus any insertion loss of the attenuator. A block diagram of the connections is shown in Figure 2.

A number of close-spaced radio relays in the 160 megacycle band were measured during the past summer by the method outlined. Fifty-watt base station type repeaters of one manufacturer were found to require between 55 and 56 db attenuation between transmitter and receiver. Three to five watt battery type automatic repeaters were also measured and found to require between 38 and 48 db isolation. The frequency separation in all instances was approximately 1.5 megacycles.

Antenna Coupling Curves

Having determined the required attenuation between transmitter and receiver, curves can be consulted to determine the required physical spacing between the transmitting and receiving antennas. Curves giving the comparative attenuations between coplanar and colinear antennas at various spacings are shown in Figure 3. It is apparent that colinear, or vertical spacing, provides the greatest attenuation for a given distance greater than about 5 feet in the 160 megacycle band and 17 feet in the 30 to 40 megacycle band.

A Decoupling Support Stub

Methods were investigated for increasing the attenuation between antennas without increasing the physical spacing. It was found that coupling exists on both the support structure and on the RF transmission lines feeding the antennas. Such being the case, a detuning stub as described by King[2] should be effective. One was constructed from copper water pipe and found to add 20 db isolation over theoretical and measured isolation without the stub. The principal consists of inserting a high-impedance stub at a point at which a current maximum (voltage node) would exist if the stub were not present. Probing the support structure with an RF indicating device consisting of a microammeter and diode crystal rectifier verified that voltage nodes exist at points any integral number of half-waves from the center of half-wave antennas or the base of ground-plane antennas. A diagram of the constructed support structure is shown in Figure 4. It was found that the isolation between antennas was not only increased 20 db but that the standing waves were removed from the outside of the transmission lines to the extent that coupling by this method was no longer a problem. Placing a decoupling ground plane on the support structure added another 6 db to the attenuation but adding a second ground plane on the opposite antenna apparently nullified the effects of the first.

As constructed, the support structure presented a number of mounting problems. It was not self-supporting to the extent that it could be expected to withstand winter wind and ice loading conditions. When mounted on a metal lookout tower, there was a 5 db reduction in effective attenuation. Pole mounting was satisfactory if cross-arms were used to support the antennas. It is necessary to adjust the shorting bar after the antenna is in place. This adjustment is rather critical. It can best be performed by attaching an RF voltmeter to the receiver transmission line and adjusting the shorting bar for null. See Figure 5.

Measuring Antenna Isolation

The actual attenuation between any two antennas may be conveniently measured

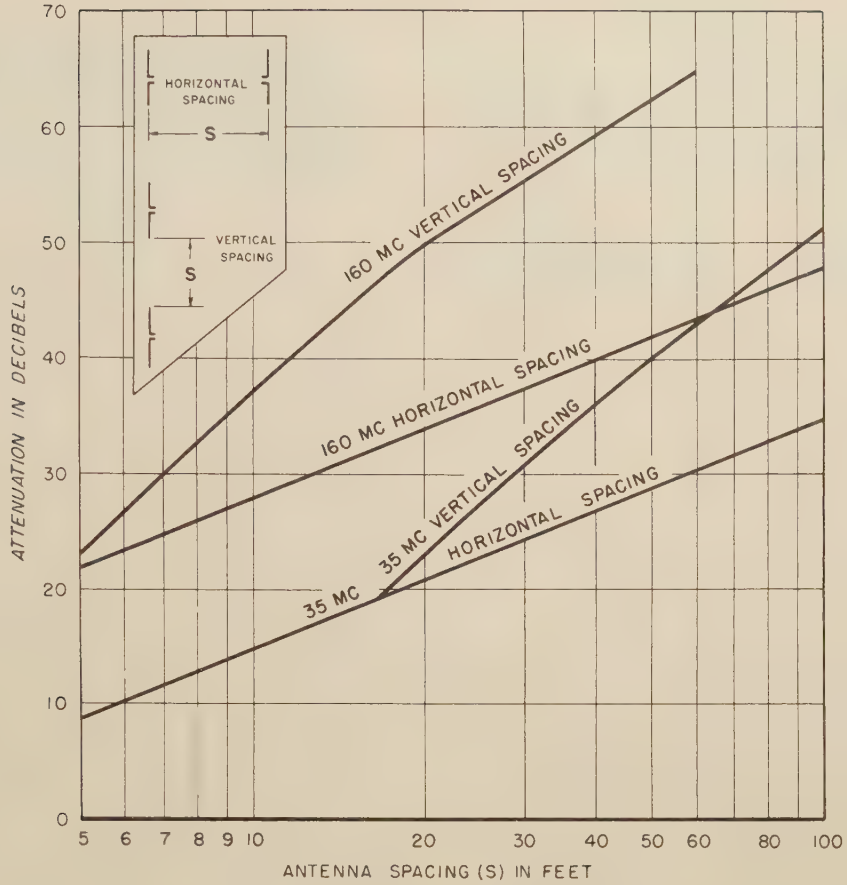

Fig. 3 - Antenna attenuation.

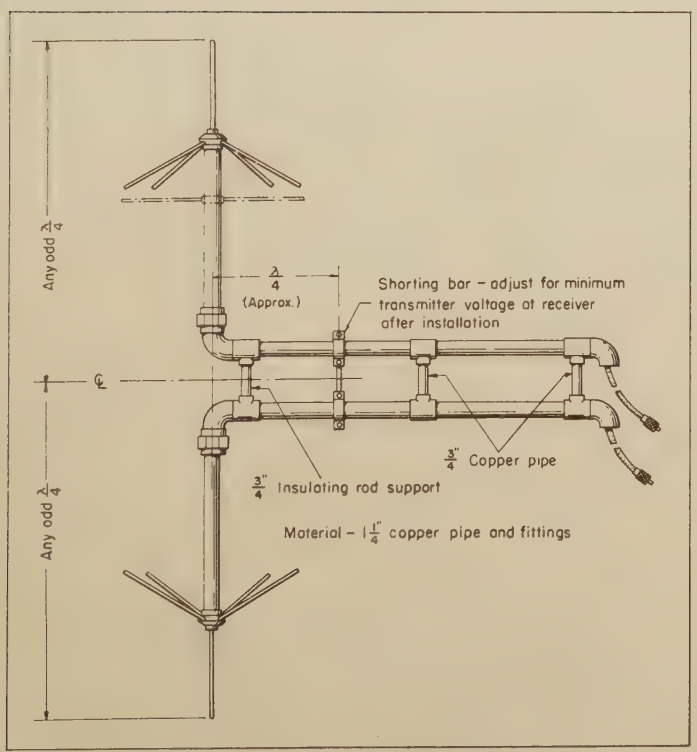
Any odd λ/4
λ/4 (Approx.)
Shorting bar - adjust for minimum transmitter voltage at receiver after installation
℄
Any odd λ/4
3/4" Copper pipe
3/4" Insulating rod support
Material - 1 1/4" copper pipe and fittings

Fig. 4 - Decoupling antenna support.

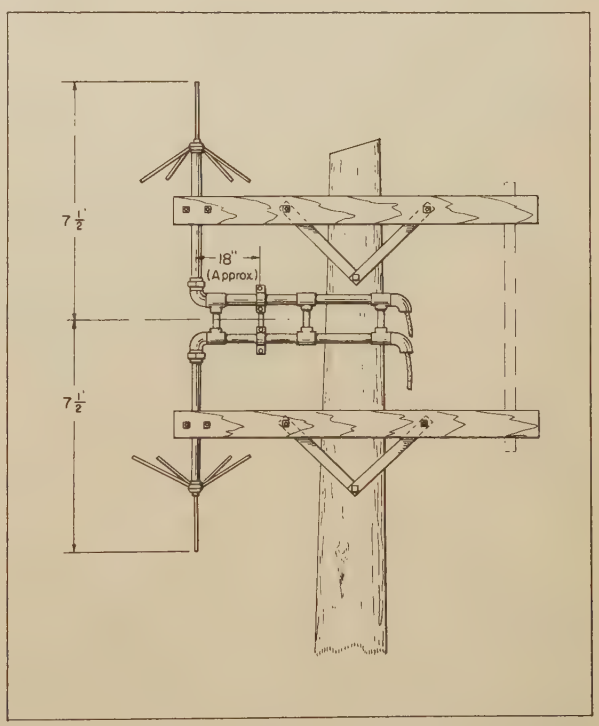

Fig. 5 - Practical mounting arrangement, with supporting cross-arms.

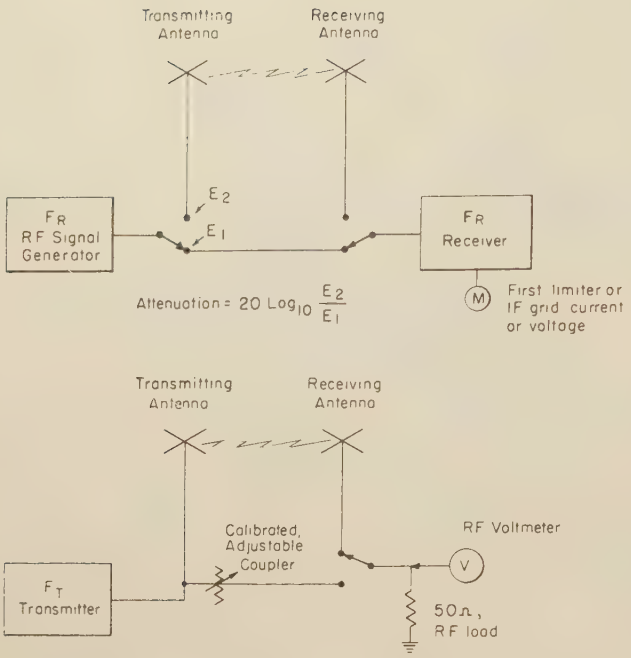

Fig. 6 - Methods of measuring attenuation between antennas.

by two methods. The receiver may be attached to the receiving antenna and used as an RF voltmeter. The signal generator is first attached to the receiver and a reference established which is well above any local noise level. The signal generator is then attached to the transmitting antenna and the output increased until the same reference is again obtained. The ratio between the two signal generator readings in decibels is the measured attenuation between the two antennas.

The second method consists of using the transmitter as an RF source and connecting a radio-frequency voltmeter to the receiving antenna. The transmitter is then connected to the RF voltmeter via a variable attenuator which is adjusted until the voltage reading is the same. The attenuation via transmission lines and antennas is equivalent to that read on the attenuator if the insertion loss is also included. Block diagrams of the methods are shown on Figure 6.

Desensitization Measurements

Any actual desensitization that may exist at a particular installation may be measured by connecting a signal generator to a third antenna placed outside the immediate field of the normal transmitting and receiving antennas and increasing the output until the threshold level is established, either 12 db SINAD or 20 db noise quieting. The transmitter is then turned on and any increase in signal generator output noted that might be required to again establish the threshold reference level. The ratio of signal generator reading #2 to reading #1 in decibels is a measure of the desensitization. See Figure 7.

Conclusion

There exist wide differences of opinion as to necessary antenna spacing for a given pair of equipment to accomplish satisfactory simultaneous transmission and reception. It is apparent that some of the contributing factors have not been fully recognized. Probably the most significant is the local noise level. Theoretical considerations of receiver sensitivity are based on noise free locations, i.e., bench sensitivity. For truly good receiver front ends, this is ordinarily not realizable and receiver sensitivity may be reduced a few decibels to as much as 20 or more decibels in severe cases depending upon location and local noise level. This means that desensitization by the associated transmitter can be obscured. It has also been determined that couplings exist between the transmitter and receiver by means of other than the respective antennas. Under certain conditions the effects of one type of coupling may nullify, at least in part, the effects of another. This has been demonstrated in the laboratory by

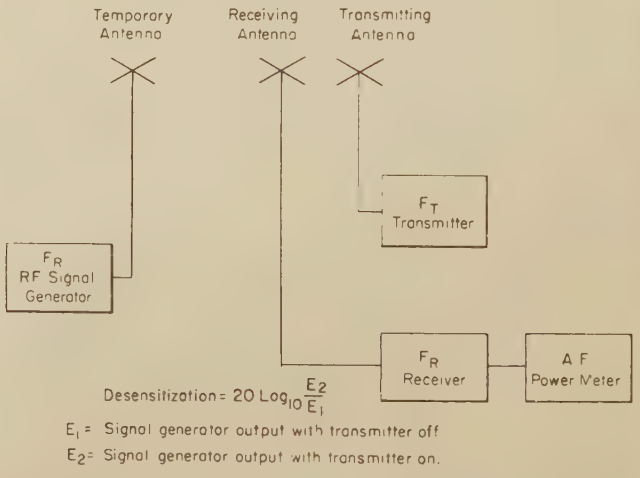

Fig. 7 - Method of measuring desensitization on location.

coupling energy directly from the transmitter to the receiver in addition to the antenna coupling by means of a variable attenuator and, by controlling both the amplitude and phase of the coupled energy, a high degree of neutralizing may be accomplished. It is also known that structures, such as metal

lookout cabs, interposed between the two antennas decreases and in some cases may increase the coupling. There are probably other factors that enter into producing a satisfactory operating repeater system; however, determining the required isolation, measuring the actual isolation, and then measuring any incidental desensitization will provide a practical evaluation for the necessary antenna spacing.

Bibliography

[1]For a more accurate description of transmitter noise spectrum see "The Gaussian Curse-Transmitter Noise Limits Spectrum Utilization" by N. H. Shepherd and J. S. Smith, TRANSACTIONS IRE-PGVC, April 1958.

[2]Antennas Transmission Lines and Wave Guides by King, Mimno, and Wing, p. 151.

A REPORT ON INTERFERENCE CAUSED BY INTERMODULATION PRODUCTS GENERATED IN OR NEAR LAND MOBILE TRANSMITTERS

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Summary

This report is a condensation of published papers combined with unpublished information on the subject of intermodulation products generation and interference created in or near land mobile transmitters. The following papers have been used in preparation of this report:

Radio Report No. 50, "Investigation of Intermodulation Interference in the 100-108 Mc/S Mobile Radio - Telephone Service". Report prepared by Mr. C. H. Sturton assisted by Messrs. G. J.Campbell and B. T. Harnett of the New Zealand Post Office.

"Interference Due to Intermodulation Products UHF Maritime Service", Special Committee No. 36 of the RTCM.

N. H. Shepherd, "BI-Dimensional Interference Analysis", Proceedings of the Third Conference on Radio Interference Reduction.

Each of these papers provide excellent material that is useful as a continual reminder of the effects of interference and practical techniques for tracking down and reducing interference. A particular type of interference, intermodulation, is troublesome to locate since it can be generated in a variety of devices.

Introduction

Intermodulation products are generated when two or more frequencies mix in a non-linear element. When intermodulation occurs, an infinite number of new frequencies are generated and each is given a number according to its order. For two frequencies A and B, the order of the products generated are as follows:

Second Order: $A \pm B$,

Third Order: $2A \pm B$, $A \pm 2B$,

Fourth Order: $A \pm 3B$, $2A \pm 2B$, $3A \pm B$

Fifth Order: $A \pm 4B$, $2A \pm 3B$, $3A \pm 2B$, $4A \pm B$

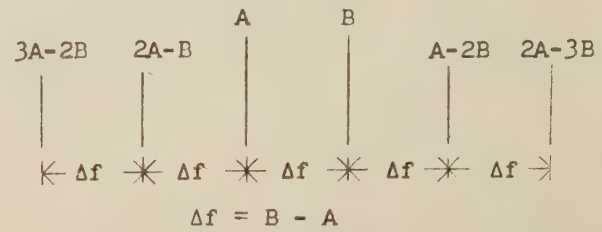
Nth Order: $A \pm (N-1)B$, $2A \pm (N-2)B$---$(N-1)\,A \pm B$

Only a small number of these products are significant in creating interference. The most important ones are the odd order products which fall near the frequencies A and B as follows:

Third Order: 2A - B, A - 2B

Fifth Order: 2A - 3B, 3A - 2B

These odd order products are distributed around frequencies A and B as follows:



Findings of the New Zealand Post Office

The New Zealand Post Office, at their Brooklyn base station, operate on the following frequencies:

75.5 Mc
101 Mc
102.5 Mc
103.95 Mc
105.45 Mc
106.85 Mc
107.5 Mc

To prevent intermodulation in transmitters and receivers, it was necessary to install high "Q" cavities in the coax line of equipment operating on each of these frequencies. The cavity filters were effective in eliminating interference generated in the transmitters and receivers, however, it was found that intermodulation interference was not completely eliminated.

It was suspected that intermodulation was occurring in some non-linear conductor external to the equipment and to varify the condition, a small germanium rectifier was connected to a dipole. When this device was outside near the antennas, the interference was reproduced.

A systematic search was then started to track down each non-linear connection around the antenna structure. To facilitate the location of non-linear conducting elements, a search head, see Fig. 1, was used. When brought close to a non-linear conducting element, the rectifier therein produced a modulated difference frequency which is picked up by the receiver connected to the center loop. The following points were found to be frequent sources of rectification:

1. Between 3-bolt clamps and stay wires.
2. Between stay wires and thimbles.
3. Between thimbles and shackle pin.
4. Between shackle pin and shackle.
5. At threads of rigging screw and at rigging screw tie wire.
6. Between shackle pin and lower shackle.
7. Between shackle pin and anchor rod.

On one occasion, it was found that 14-year old American truck parked 90 feet from the antenna towers was a prolific source of intermodulation interference. The interference was eliminated by driving the truck 200 feet from the antenna towers.

In order to obtain quantative data on the distance at which intermodulation interference created by a non-linear joint would be troublesome, a series of tests were made using the setup in Figure 2. Results of this test are shown on Figure 3.

In all instances of intermodulation arising from rectification in loose metallic joints, it was found possible to remove the interference by either insulating the metal-to-metal joints or by cleaning and bonding the joints together.

Report of RTCM SC-36

Intermodulation products on board vessels were investigated by Special Committee No. 36 of the RTCM. The most serious intermodulation interference was caused by television transmitter signals combining with the signal from the VHF transmitter on board. The VHF transmitter and receiver operate duplex and are separated by the same frequency difference, 4.5 Mc, as the video and sound carrier of the TV station. The three transmitter frequencies, VHF, Video, and Sound, combine in non-linear elements to produce a third order product equal to the receiver frequency.

Measurements were conducted on board the U.S.C.G. Cutter FIR and the U.S.W.S. FREDERICK FUSTON. The results of these tests are shown in Figure 4.

Transmitter Intermodulation

Intermodulation products generated in transmitters can be measured by using the test setup shown in Figure 5. An external signal is introduced into the transmitter under test. The cavity filter tuned to frequency f_o is used to prevent intermodulation in the receiver.

Typical levels of intermodulation products created in transmitters are shown in Figure 6. Each curve shows the attenuation of the intermodulation product below the level of the signal introduced into the transmitter under test.

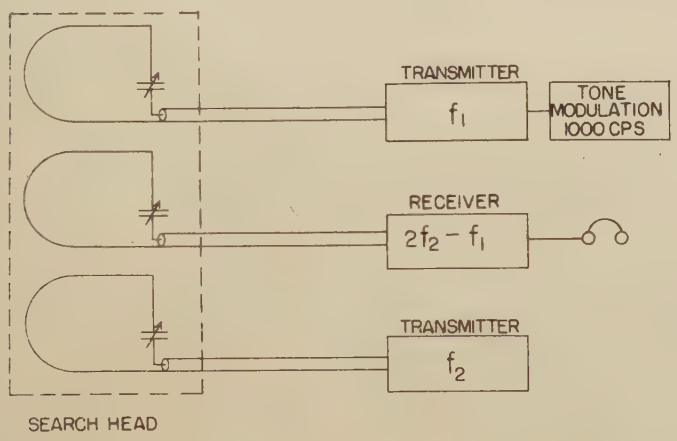


Fig. 1 - Search equipment for locating nonlinear joints.

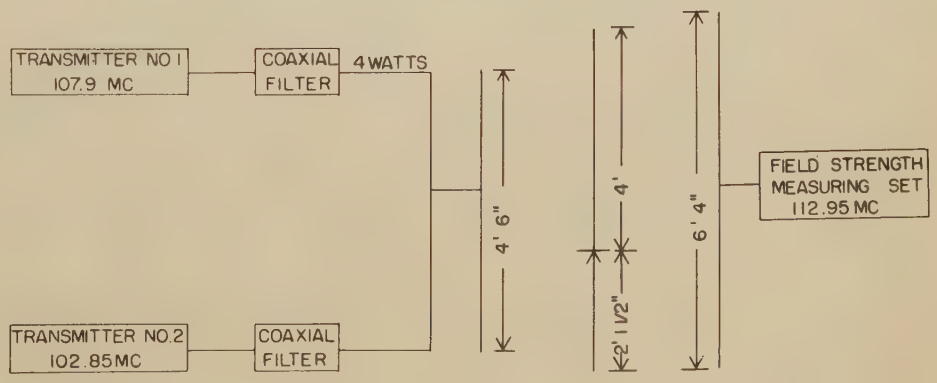


Fig. 2 - Test setup to measure intermodulation radiation from nonlinear conductors.

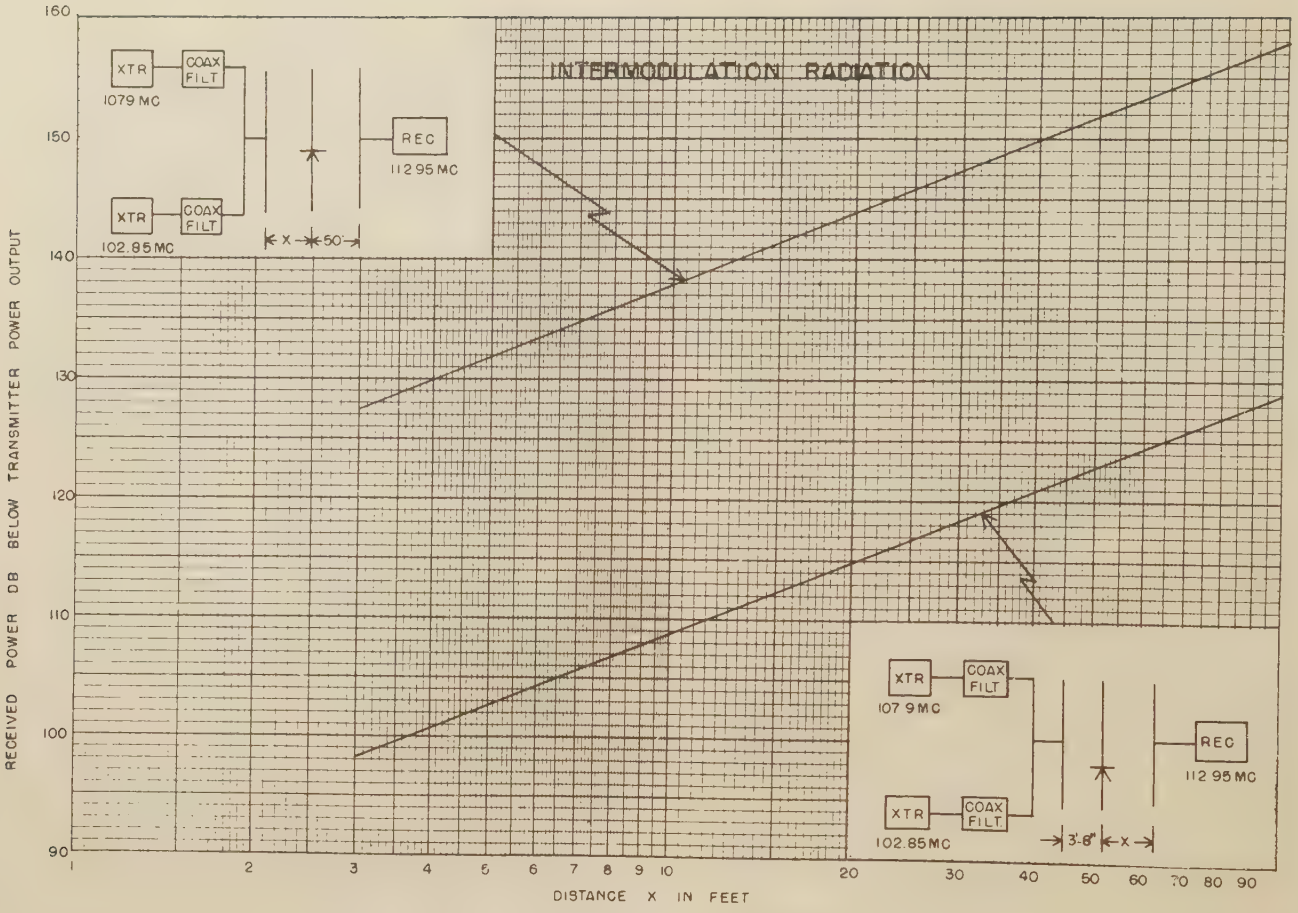


Fig. 3 - Intermodulation radiation.

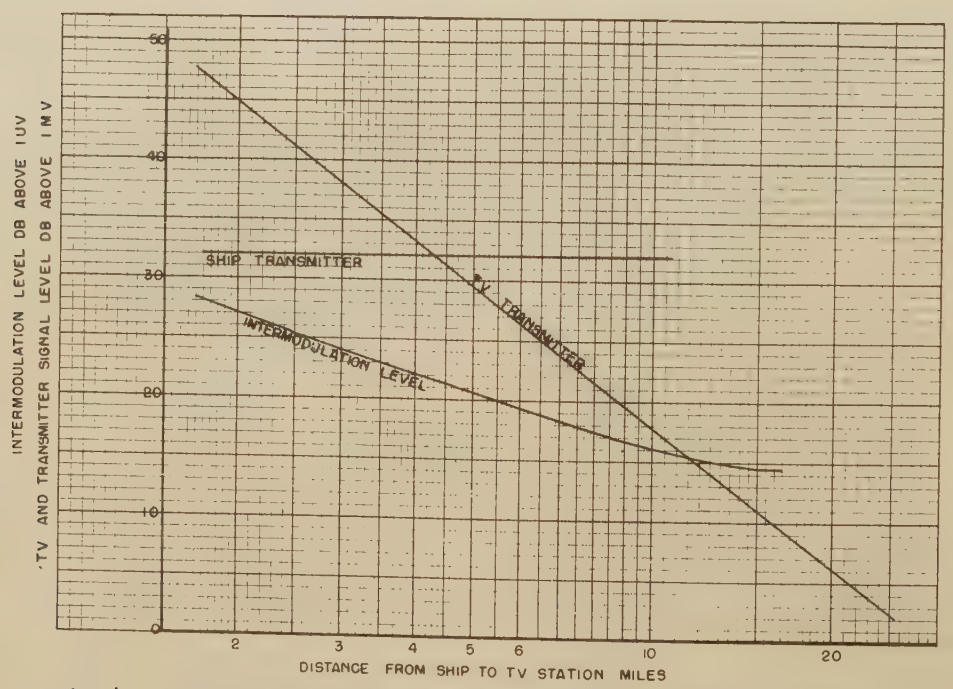

INTERMODULATION LEVEL DB ABOVE 1UV
TV AND TRANSMITTER SIGNAL LEVEL DB ABOVE 1MV
SHIP TRANSMITTER
TV TRANSMITTER
INTERMODULATION LEVEL
DISTANCE FROM SHIP TO TV STATION MILES

Fig. 4 - Intermodulation level on board U.S.C.G.C. FIR; 4.5 mc duplex separation.

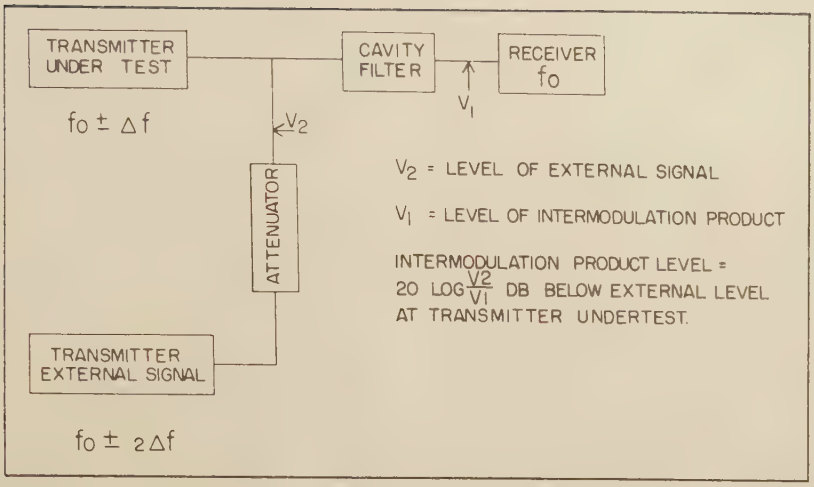


Fig. 5 - Test setup for measuring transmitter intermodulation.

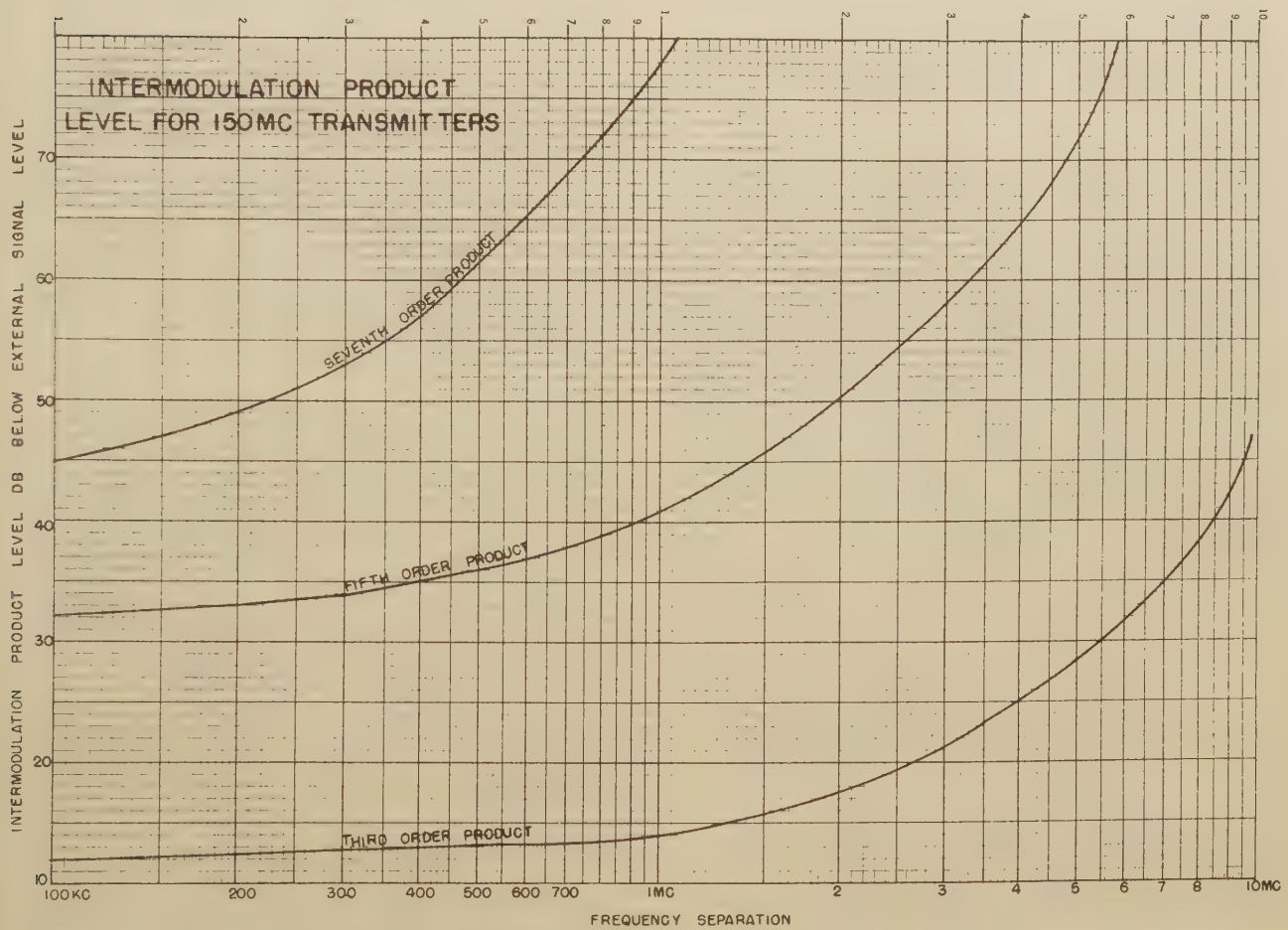


Fig. 6 - Intermodulation product level for 150 mc transmitters.

A VHF POCKET RECEIVER*

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It was not too long ago that high quality FM two-way radio was introduced in the mobile field. From the first it was recognized that a portable two-way communications package would be desirable. The first portable units gained real reknown during World War II as the Walkie-Talkie. With this design experience as a base, Motorola introduced the first "Handie-Talkie Radiophone" in 1948. It performed a yeoman job in my vital services and excited the imagination of many who could see the need for personal communication. Since that time, the transistor has come along and provided a means to reduce the size and weight, and increase the reliability of portable equipment. In 1956, Motorola introduced their transistorized line of two-way Handie-Talkie equipment.

In the field of personalized pocket carried equipment, the units remained limited to super-regenerative tube sets. The transistor, with its promise of extreme reliability, low drain, and small size, opened up new possibilities in this field. When the high frequency transistors became available, Motorola went to work. The results: (See Figure 1) a full-fledged high quality FM set that is really portable. The "Handie-Talkie Pocket Receiver" introduced in late 1958 is a double super-hetrodyne receiver with two separate crystal oscillators. It has 16 transistors and 6 diodes to produce results of a full size mobile unit. (See Figure 2.) For example:

1) Sensitivity 1 uv. at 150 MC and .5 uv. in the 50 MC band.

2) Weight: 10 oz. with a rechargeable sealed nickel cadmium battery or 12 oz. with a mercury battery that will last up to 200 hours.

3) 100 milliwatts of audio power.

4) Selectivity of 45 db on the adjacent channel.

5) Squelch sensitivity of .5 uv. in the 150 MC band or .25 uv. in the 50 MC band.

Note the unit is packaged in a high impact plastic case with a 1-3/4" speaker for real audio power output. It has an external squelch control, separate battery compartment (See Figure 3) which is accessible without getting into the radio section, and either a mercury battery or a rechargeable battery can be used interchangeably. The charging jack for the rechargeable is located in the bottom end of the case so that the rechargeable battery can be charged in the case. The case is not only high impact, but all controls, plugs, and switches are recessed to protect them when the unit is dropped. A very sturdy clip is provided to fasten the unit to a belt or pocket.

Now let us look at a few of the design factors. The design of a pocket receiver involves some interesting compromises. Producing 1 uv. at 150 MC or .5 uv. at 50 MC with a transistorized receiver is interesting, but several compromises are required to put this receiver in a 10 oz. package. The size must be reduced without impairing first of all the sensitivity since at best, a personalized receiver antenna must be degraded from a standard di-pole.

It is worthy to note that in the Bell Laboratory's Record of January, 1958†, a conclusion was reached that "To achieve this selectivity and stability, crystal-controlled super-heterodyne rather than super-regenerative circuitry will be required". A super-heterodyne receiver is required to produce adequate selectivity and stability to operate in a spectrum that is ever increasing in number of channels. Experience with the Handie-Talkie Radio Pager, a 60 to 140 KC tuned radio frequency selectively called receiver introduced in 1956 for in-plant use, indicated some of the limits in size, weight, battery life, audio output, and durability that would be required.

As a first step in the design the lowest drain, full sensitivity receiver was built. It included 16 transistors and 6 diodes. It included a high IF at 12 MC for proper reduction of the image and other spurious responses and a second IF at 455 KC. The lumped selectivity approach with a Perma Kay filter was

*Presented at a meeting of the Chicago Chapter of the PGVC in January, 1959.

†Bell Laboratories Record, "Pocket-Radio Signalling", by W. Strack, January, 1958.

required in the 455 KC IF for adequate selectivity. The 455 KC IF is terminated in a limiter and discriminator for detection of the FM modulation. This is followed by an audio driver transistor which drives a push-pull audio output stage to 100 MW. The squelch circuit includes both a noise amplifier and rectifiers to cut off the audio for the dual purpose of noise cancellation and audio stage drain reduction. The first and second conversion are accomplished in diode mixers which requires no drain, yet produces a sensitivity well under 1 uv. at 150 MC. The second mixer is energized by a transistor oscillator. The first oscillator is multiplied four times in a diode multiplier which again requires no drain. In the 150 MC band, the first mixer is preceded by three coils of RF selectivity, while in the 50 MC band, an RF stage is added to reduce the sensitivity to better than a guaranteed maximum of .5 uv.

The choice of transistors and their operating points was always made with drain in mind so that the resulting total receiver drain when squelch is under 5 ma. at 10 volts. To accomplish this, the audio is practically cut off and most IF amplifiers operate as low as 0.2 ma. at 3 volts. This, of course, is very conservatively under the transistor rating. All transistors are pre-aged for long life and reliability. At the 100 MW audio output level, the drain is 35 ma. at 10 volts for a total efficiency of over 30%.

Next, a package size which was compatible with the market requirements was selected. The housing was attractively styled in a high impact plastic. The plugs and controls were recessed for protection and a rugged clip added so the unit could be secured to a belt or pocket. A 1-3/4" speaker was selected to take maximum advantage of the 100 MW of audio output considered necessary in this type unit. A battery pack that would power the unit for up to 200 hours consisting of commonly available mercury cells was deemed a necessity. The case was designed with a separate battery compartment for battery change by personnel who should not have access to the radio. The battery compartment was also designed to accept a rechargeable nickel cadmium battery. This is a sealed cell which will have many years of life. Over 60% of the final weight is now included.

Now with an antenna added and the radio built into its section of the case, the design is complete. The best antenna arrangement derived resulted in a real space saving. It consisted of a di-pole with the radio chassis as one section and the batteries acting as the other.

To house the radio, a chassis was constructed by molding a cellular structure of plastic and gold plating this structure for shielding. The individual stages of the radio were then constructed on separate printed boards to form modules and inserted into the chassis. (See Figure 4.) This provides a very rugged structure which could be contoured to clear the large 1-3/4" speaker and its magnet. The radio stages were also completely accessible for servicing, with every connection in the radio appearing at the base printed boards. This modularized construction also means that each stage in the radio can be removed separately by disconnecting its input, output, power, and ground connections. This basic radio chassis was then floated in the radio housing on sponge rubber to absorb the shock encountered when the unit is dropped.

The receiver has several accessories such as a lapel speaker used to bring the voice closer to the ear in a noisy location. An ear piece and volume control is included for areas of even higher noise level. The 50 antenna termination is also brought out so that the unit could be attached to an antenna in a vehicle or any desirable improved antenna.

At this point, the question always arises as to how the unit can talk back. For this requirement, a companion transmitter with 1 watt at 150 MC and 1.25 watts at 50 MC has been introduced. (See Figure 5.) It is housed in a package weighing 44 ounces. Its antenna is in the microphone cord. The inside of the unit is shown on Figure 6.

The combination can be carried on the belt or in an attractive leather carry case on a leather belt. (See Figure 7.)

The uses for this type communication are many. For example, the maintenance man, the plant guard, or a police officer can use this equipment very effectively. (See Figure 8.)

The transistor has made the equipment possible, with its small size and low drain. But, with its performance and reliability, it has opened up a new era in personal communications.

Fig. 1

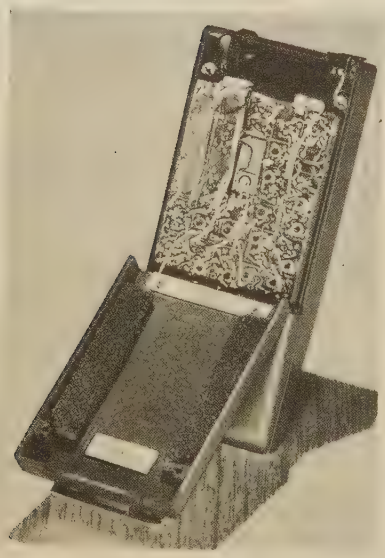

Fig. 2

Fig. 3

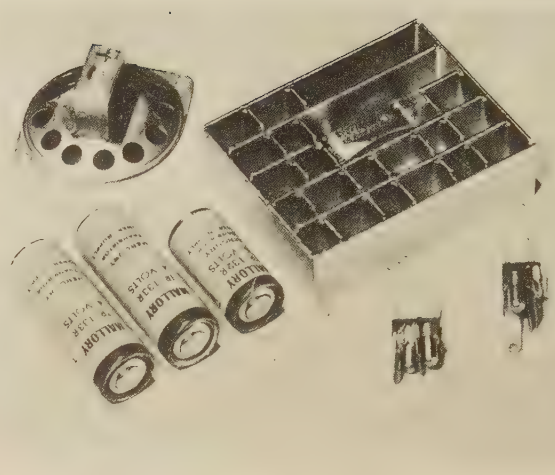

Fig. 4

Fig. 5

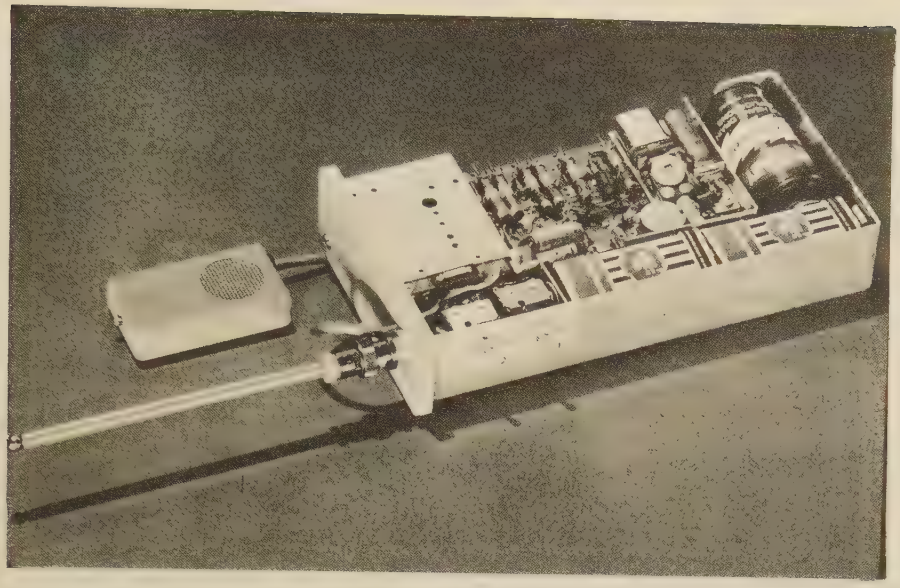

Fig. 6

Fig. 7

Fig. 8

MOTOR CARRIER MOBILE RADIO SYSTEMS*

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Why Do Motor Carriers Purchase Radio Equipment?

1. To lower operating costs.
2. To improve operating efficiency.
3. To improve customer relations.

Operating costs may be lowered by decreasing idle time of vehicles and drivers - decreasing back-track mileage - virtual elimination of telephone calls by drivers, charges and waste driver time.

Operating efficiency may be improved by the direct supervision of drivers allowed by radio - close contact between drivers at the scene, and the OS & D and collection personnel at the terminal - the quickly available information between the city driver picking up the freight, and the terminal personnel who must provide for its over the road movement.

Customer relations are greatly enhanced by the use of radio - fewer pick-up requests have to be carried over to the next day - hot shipments may be quickly picked up throughout the day - consignees may be given accurate information on delivery of incoming shipments - problems of shortage, damage, payment, etc. can be settled on the spot via radio.

Advantages of a Properly Designed and Installed Radio System

Most trucking people know little about radio - we depend upon the manufacturers to engineer and build suitable equipment - we depend upon their sales personnel to present fairly to us what the equipment can do - we don't want to be over sold or under sold.

In each city in which we operate the situation is usually somewhat different - these may be operational differences requiring greater or lesser area of coverage - terrain differences - interference from industrial and other sources - frequency crowding, etc.

It is important to the trucker to purchase the system, whether it be 44 mc, 150 mc, or 450 mc, that best suits his needs - cost, coverage, and fidelity of transmission are factors in this selection.

Antenna location is important - can it be located on terminal premises and if so how high should the tower be - if on a rented site, what is the best location for necessary coverage and lowest rental also least land-line charges - if the antenna is located on or near a lake front, is it a directional antenna.

The central point must be properly located, and attention should be given to running hand sets off the control points, or locating additional control points.

We are often unaware of additional benefits we may derive from radio - the radio salesman is sometimes unaware of needs we have which could conveniently be met often by slight additions to the system.

A thorough discussion and understanding between the motor carrier purchaser and the radio seller should always be a prerequisite of purchase.

Channel interference, crowding and other interferences effecting carrier usage.

Radio equipment must be designed to eliminate as much as possible frequency straying - good maintenance of radio equipment is an important factor in preventing this straying. Stray reception from other systems can be source of confusion to our dispatchers and drivers - radio systems in which the receivers are opened by a transmitter tone have done much to eliminate this, as well as prove helpful in frequency sharing.

Frequency sharing can be done satisfactorily, but can prove very unsatisfactory. Cooperation between motor carriers, the radio industry and the FCC, I hope, will tend to decrease the need for such sharing. Advance radio design, I have no doubt, will make it less necessary.

We must call upon the radio industry to perfect radios that will be less subject to dead spots, skip, and industrial noise. In the few years I have dealt with motor carrier radio it seems to me you have done a very good job in this direction.

*Presented at a meeting of the Chicago Chapter of the PGVC in May, 1959.

Size, Design and Installation of Radio Mobiles

Tractor cabs are becoming more compact, have more equipment in them, with consequent less room for radio equipment.

Chassis housing must be strong to prevent damage - chassis must be cushioned to prevent shock - even so when mounted on the floor under the dash, or back of the seat it is subject to blows, receives great vibration and must often be removed for maintenance and repair on the tractor - I believe a flat unit hung from the cab roof would be out of the way, and receive less damage from vibration and blows.

Chassis should be easily removeable from housing so that defective units may be replaced and vehicle put back into service as quickly as possible.

Leads from mike, speaker and antenna should be carefully installed so they would not have to be torn out for vehicle repair.

Tube units should be wired through the vehicle switch to prevent radio being left on overnight. This has the disadvantage of the driver not being able to receive if he is parked at a dock with engine switched off.

Transistorized equipment seems to provide good answers to needs for more compact equipment, equipment less subject to damage, and low power drain.

With much of the equipment in use today the radio salesman should go into the subject of heavy duty batteries and low cut in generators with the motor carriers purchaser.

Mobile unit antennas should be as small as possible as they are vulnerable to damage - I believe they should, if possible, be of a material that will spring back to original position after being bent, as I understand a bent antenna will cause poor reception from the reverse direction of the bend.

Radio mikes, in my opinion, should have handles - almost all drivers wear gloves, and a handle would be much easier to grasp and would eliminate the glove muffling the mike - in most radio procedure instructions seem to be to talk "across" the mike - most drivers don't - why can't mikes be built so that when driver thinks he is talking into it, he is actually talking across its face.

The speaker should be designed so that it can be, and it should be mounted as nearly on level with the drivers ear as possible - preferable position seems to me to be left hand upper front corner of the cab, or just above the left door.

Presently Possible, or Feasible future Motor Carrier Radio Uses Other Than Pick-up and Delivery Mobile

At the present time we make use of mobile units in our city wrecker and service vehicles - saves time and backtracking of these units gets out of service of P & D vehicles in service faster.

In several locations we use radios in our spotter vehicles within the terminal yard gets trailers in and out of dock doors faster.

Our city safety and fleet supervisors use mobile units, which make them easily located and speedily available for on the spot intervention in problems.

In some of our larger docks the foreman and supervisory personnel could use small hand-carried equipment to good cause.

Many truck lines have used central freight checking equipment and have removed it due in great part to the fact that it is cumbersome and highly susceptible to damage - radio equipment of the hand-carried type might eliminate the bad features of this checking system, and bring it into general use.

As pick-up areas expand, and as numerous customer locations come into being between cities such as Chicago and Milwaukee, automatic radio relay stations may become very vital to our operation.

On the Buck Rogers side what are the possibilities of pick up units signalling to the terminal, and being automatically given pick ups from a machine such as Univac or Ramac which has taken pick up calls, vehicle locations, empty capacity of units, etc., has sorted and matched these factors and has made the proper distribution.

The use of microwave presents a tremendous area of useful possibilities for motor carriers, one minor but interesting one that occurs to me, is the recording and accounting for gasoline placed in our units (approximately 20,000 gallons per day) a telemetering could be installed relayed by microwave pumps would be activated by two plates similar to charge-o-plates - one for the vehicle, the other for the person authorized to put gas into the vehicle - almost instantaneously our general office in La Crosse would have available a

record of the unit drawing the gas, number of gallons drawn, what city it was in, who put the gas in, and, if it desired, the remaining gallonage in the tank.

[This article presents a good picture of what the users in the motor carrier radio service expect from two-way radio. Although not very technical, sections of the paper provide some thought-provoking ideas. -- Editor]

AN ANALYSIS OF RADIO FLUTTER IN FUTURE COMMUNICATIONS

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Summary

It is expected that future military operations will be extremely mobile. It is planned to provide for the Army full duplex communications while its subscribers are in motion. This will require the use of radio for serving the subscribers.

Reception from or in a vehicle in motion is plagued by a malady known as flutter. It results in cyclical impairment of reception. It is caused by the varying field strength pattern generated on multiple path reception of the same radio signal. The audio signal level can be maintained by an AVC. The increase of noise level during operation of AVC can adversely affect the operation of system signaling.

All sorts of schemes are being studied to overcome the effect of the flutter. In this paper the flutter effects on the communication system will be analyzed. The improvements resulting from certain remedies will be predicted. Propagation data at 150 mc will be presented. Emphasis will be placed on the switched antenna type of space diversity reception. The value of this technique will be established by a brief discussion of the benefits that may be expected.

Recent advances in the Solid State Field make feasible the design and construction of an applique unit which will assure us of benefits of space diversity reception at a relatively low cost. The design of antenna switching circuitry which operates on a programmed basis will be described. Problems generated by these techniques will be aired.

Introduction

The requirements of our civilization for the ability to communicate while in transit for both military and business operations are rapidly increasing. This immediately implies that at least a part of the transmission linkage be free of the restriction imposed by wires and consequently demands a much greater use of radio for carrying the message. Radio reception while in motion becomes a luxury which can be enjoyed if the price is paid.

Reception from, or in a vehicle in motion, is adversely affected by a phenomenon known as flutter. This phenomenon results in cyclical impairment of reception. It is caused by the varying field strength pattern generated by multiple path reception of the same radio signal. The audio signal level can be maintained by an AVC. However, the increase of noise level, due to the functioning of the AVC, can put limits on the effectiveness of system signaling. The reduced signal to noise ratio may not impair intelligibility but will degrade the quality of voice transmission. A study of the phenomenon of flutter and recommendations to cancel or reduce its effect are the objectives of this paper.

The Radio System

A communication system having requirements for radio subscriber's loops up to 10 miles long will be assumed in the following discussion. Two conditions of operation will be considered. In each case the transmission frequency will be 150 megacycles, the transmission power 11 watts and the receiving antenna of $\frac{1}{4}$ wave design mounted 6 ft above the ground. The transmitting antenna will be of $\frac{1}{2}$ wave design. In one mode of operation this antenna will be 6 ft above the ground and in the other 25 ft above the ground. It is believed that these conditions as stated, will represent characteristics of two significant types of operation likely to be met in the field. The first, using the 6 ft transmitting antenna represents the condition in which the radio central is operating on a temporary basis and there was not time to erect the high antenna. The other represents the operating condition in which there has been time to erect the more efficient higher antenna. A Field Strength Study[1], was made by Messrs. Charles F. Meyer and Don Soule of Motorola, Inc. on comparable radio installations. Data has been freely borrowed from this study in the preparation of the charts accompanying this paper.

The distribution curve of Figure 1 shows the probability of various loop lengths occuring in a 10 mile system. The referenced Motorola report contains information from which curves showing average field strength versus distance from transmitter can be plotted. This has been done in Figure 2. Curve A represents the condition of temporary operation with the low antenna and Curve B represents the condition of operation with the high antenna. Figures 1 and 2 have been combined in Figure 3 to plot the average field strength at the radio subscriber site for normal loop length distribution. On this sheet as in Figure 2 curve A represents a

half-wave transmitter antenna height of 6 ft and curve B represents a half-wave transmitter antenna height of 25 ft. Measurements were made with a quarter wave antenna 6 feet high. Line JK represents the receiver noise reference level for the conditions under which these data were obtained. Line LM represents the field strength necessary to achieve a 13 db signal to noise ratio under conditions of 100% carrier modulation. In the signaling system assumed each tone will be transmitted at a level of 7 db below that required for 100% modulation. The line NP represents the field strength necessary to insure a 13 db signal-to-noise ratio during signaling.

Rayleigh Distribution of Fade and Flutter

The discussion so far has dealt solely with average field strength. The term "average field strength" implies the existence of excursions and deviations from an average value. Fade and flutter are phenomena which can and will cause these deviation. The distribution and depth of fade can be predicted by the Rayleigh distribution curve. This was described by C.L. Mack in his paper.[2] Figure 4 is a Rayleigh distribution curve showing depth of fade versus probability. In the case of fading, the transmitter and receivers are fixed with respect to each other and the field strength pattern moves. This is applicable to long distance operation. In the case of flutter, the field strength pattern is likely to remain fixed and the transmitter and/or receiver will move with respect to the field strength pattern. Since these two phenomena are strikingly similar, justification for the application of the Rayleigh distribution to flutter was sought. Messrs. Norton, Vogler, Mansfield and Short listed in their paper[3] on this subject four requirements that must be satisfied if the Rayleigh distribution will apply. The conditions inherent in flutter as well as in fade satisfy these four requirements. On this basis, the Rayleigh distribution may be used for predicting the probability and depth of flutter during the operation of mobile radio subscriber loops. This distribution curve will include in its predictions the probability that the subscriber may operate in a low signal strength region even when in a fixed position during communications.

The problem of fading has been with us long enough to have received considerable thought. The result has been the conception of several remedies. One of the most favored remedies is space diversity operation. The use of dual space diversity operation provides two discrete transmission paths simultaneously. The probability of deep fades is then reduced as a square function of the original Rayleigh distribution curve. In the case of triple diversity operation the Rayleigh distribution curve is cubed. The curve of figure 5 depicts the depth of fade distribution as modified by diversity techniques. Because of the short distances involved, we can assume that the fading effects will be negligible. Hence, the discussion from this point on will be concerned with flutter. It will be treated as though it were fading.

Flutter Probabilities

The next step in the development of our objective is to combine the average field strength probability and depth of flutter probabilities. This has been done in Figure 6 by plotting field strength at radio subscriber sites for normal loop length distribution. These curves integrate the effects of flutter with and without diversity reception. The curved line is identical to curve B in Figure 3. For these data the transmitter used a $\frac{1}{2}$ wave antenna 25 ft high. The line WX represents the average field strength which would assure trouble free operation, during normal flutter 99.5% of the time. This line was plotted 22 db above the 13 db signal to noise ratio line. The 22 db figure was taken from the 0.5% point curve A-of figure 5. Similiarly the prediction of 99% trouble free operation is represented by line UV. The improvement then can be achieved by the use of dual diversity is illustrated by comparing the line QR with the line UV.

The intersection of line WX and distribution curve B indicates that under these operating conditions 15% of the subscribers can expect trouble 5% of the time. The intersection point of the line UV and the distribution curve indicate that something less than 2% of the subscribers will have trouble 1% of the time. The position of line QH with respect to the distribution curve indicates that with dual diversity reception less than .01% of the subscribers will have trouble 1% of the time.

The information plotted on Figure 7 is identical to that shown in figure 6 except that it represents performance when the $\frac{1}{2}$ wave transmitter antenna is only 6 ft high. The horizontal lines, in their relationship with the distribution curve, can predict the system operation. The intersection point of line WX and the distribution curve predicts that under these conditions of operation 85% of the subscribers may expect trouble .5% of the time. Line UV predicts

that 75% of the subscribers may expect trouble 1% of the time. The intersection of line ST with the distribution curve predicts the performance to be expected from dual diversity operation. It predicts that 15% of the subscribers will have trouble 1% of the time. The intersection of line QR with the distribution curve predicts that about 2.5% of the subscribers will have trouble 1% of the time if triple diversity reception is used.

Application of Space Diversity

In the above discussion and the accompanying charts it has been demonstrated that space diversity reception will assure more reliable performance of radio subscriber loops. Heretofore the application of space diversity has been a cumbersome and expensive proposition. In its implementation two or more antennas are required. If an antenna switching scheme were used it would involve modification of the receiver, control amplifiers and relays. More sophisticated application of the diversity concept would include the use of two or more receivers as well as multiple antennas. In this latter implementation switching is done at voice frequency and the receiving system which is favored with the batter or best reception is connected to the subscriber. Recent development has resulted in the design of a combiner circuit which functions to add, at voice frequency, all the signals that are received by the various systems. This last arrangement is the most efficient of those described.

The transistor art has reached the stage where the benefits of dual diversity can be realized at reasonable cost. Two or more antennas could be programmed alternately into one receiver. This would take advantage of the increased probability that one antenna would be in an area of high signal strength at any instant. The programming and switching and some gain can all be accomplished in a relatively simple and small transistor circuit assembly which involves no changes in the one receiver. A breadboard model of the transistorized circuit has been constructed in the laboratory. Although it was not designed for a specific radio central system the results of tests were promising.

The requirement for additional antennas, however, remains the same. The distances between receiving antennas are relatively small. The idle receiving antennas could result in distortion of the field about the active one. The transmitter antenna, during duplex operation, could result in receiver blocking. These problems could be surmounted by developing a complete antenna system for mobile use.

This paper has described a point of view from which the effects of flutter can be predicted and even minimized. The solution hinted at is not concrete. It represents one approach that may bear fruit.

Acknowledgements

The author expresses his sincere thanks to his colleagues for their heated discussions which served to spark the effort that resulted in this paper. Special acknowledgment is made for the work done by Mr. G. P. Tripp in editing the manuscript and the patience exhibited by Mrs. D. Neilson our amanuensis.

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2. C.L. Mack, "Diversity Reception in UHF Long Range Communications" Proc IRE, vol. 43, No. 10, pp. 1281-1289; Oct 1955.

3. K.A. Norton, L.E. Vogler, W.V. Mansfield, and P.J. Short," The Probability Distribution of the Amplitude of a Constant Vector Plus a Rayleigh-Distributed Vector," Proc IRE, vol. 43, No. 10, pp 1354; Oct 1955.

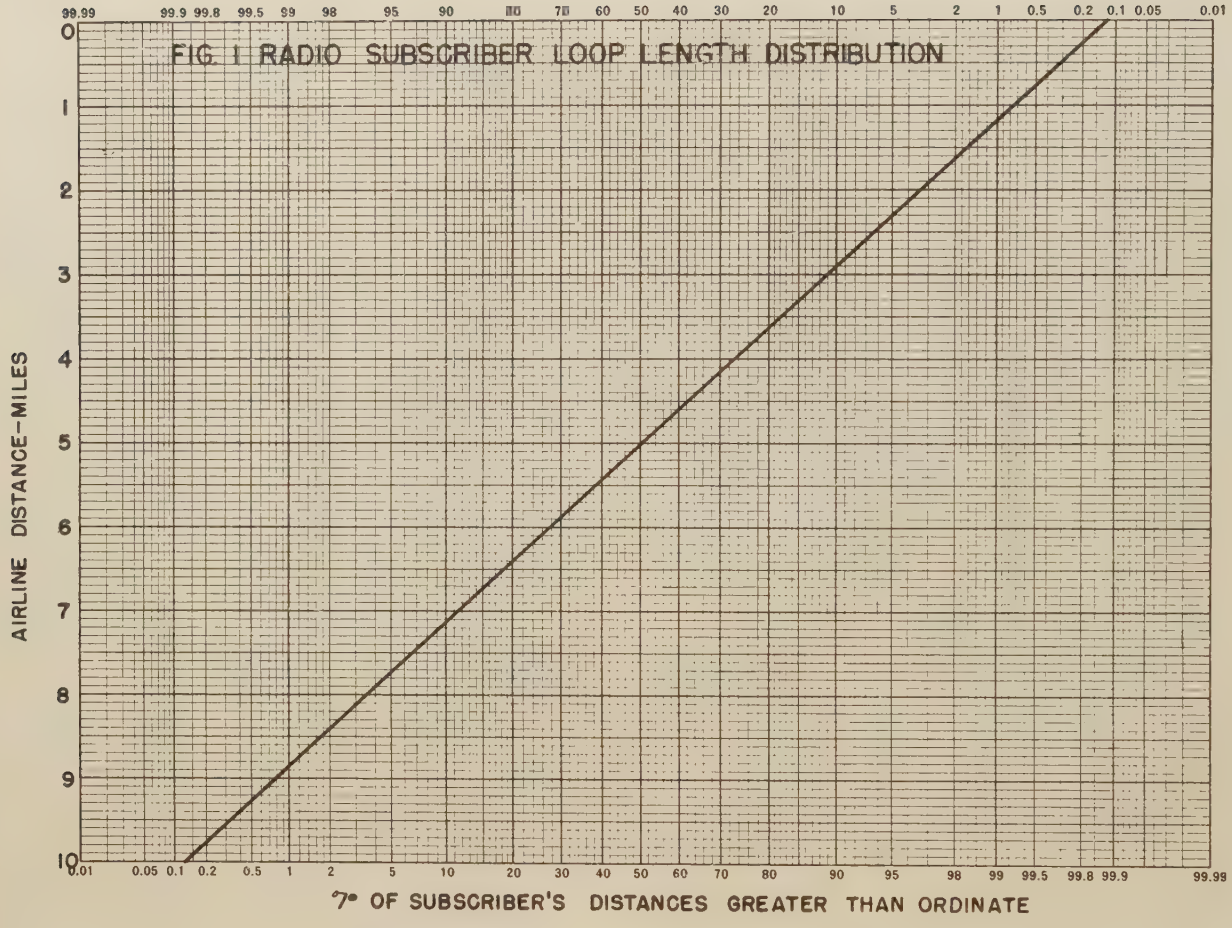
FIG. 1 RADIO SUBSCRIBER LOOP LENGTH DISTRIBUTION
AIRLINE DISTANCE-MILES
0
1
2
3
4
5
6
7
8
9
10
99.99 99.9 99.8 99.5 99 98 95 90 80 70 60 50 40 30 20 10 5 2 1 0.5 0.2 0.1 0.05 0.01
0.01 0.05 0.1 0.2 0.5 1 2 5 10 20 30 40 50 60 70 80 90 95 98 99 99.5 99.8 99.9 99.99
% OF SUBSCRIBER'S DISTANCES GREATER THAN ORDINATE

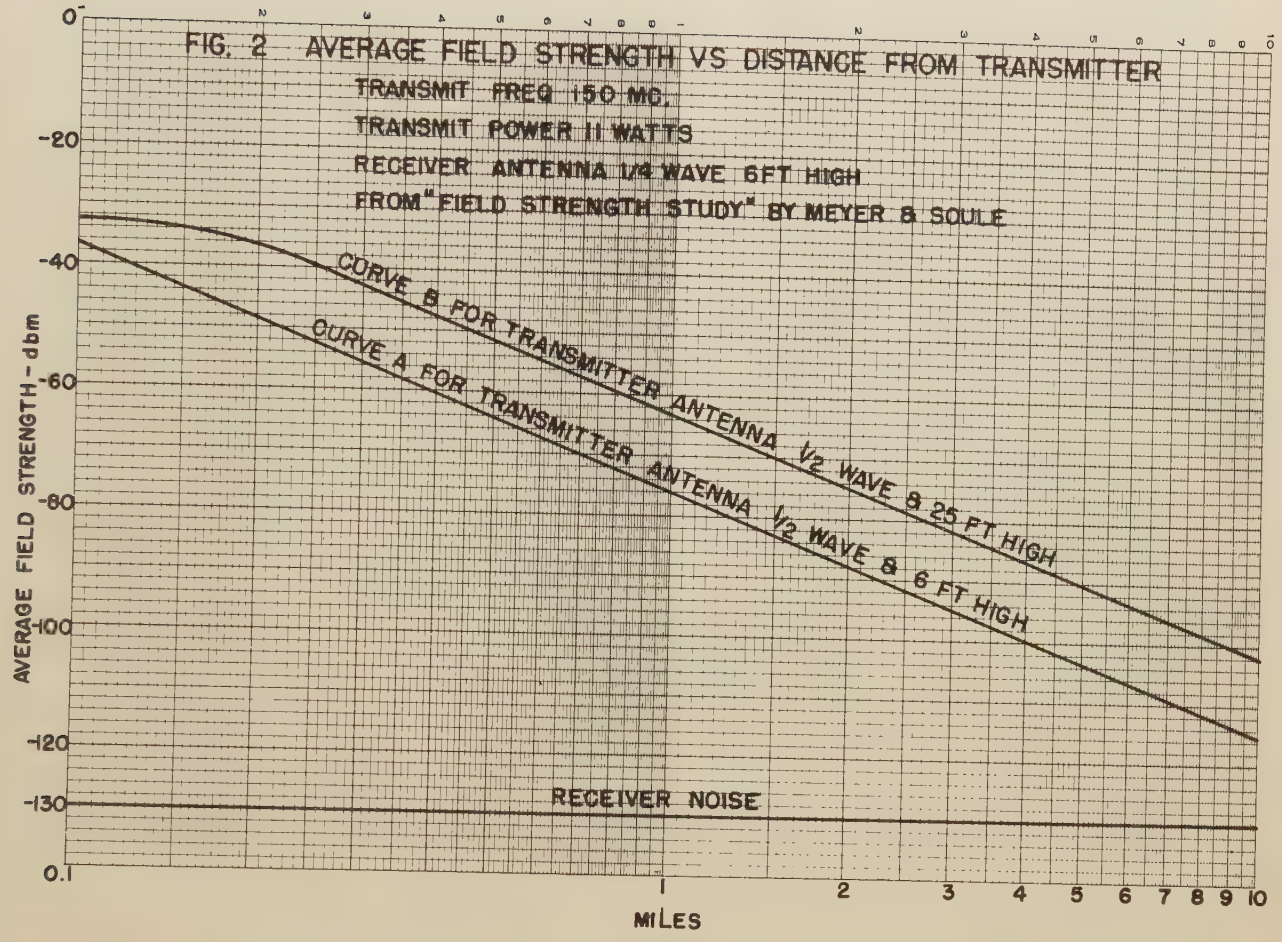
FIG. 2 AVERAGE FIELD STRENGTH VS DISTANCE FROM TRANSMITTER
TRANSMIT FREQ 150 MC.
TRANSMIT POWER 11 WATTS
RECEIVER ANTENNA 1/4 WAVE 6FT HIGH
FROM "FIELD STRENGTH STUDY" BY MEYER & SOULE
CURVE B FOR TRANSMITTER ANTENNA 1/2 WAVE @ 25 FT HIGH
CURVE A FOR TRANSMITTER ANTENNA 1/2 WAVE @ 6 FT HIGH
RECEIVER NOISE
AVERAGE FIELD STRENGTH - dbm
0
-20
-40
-60
-80
-100
-120
-130
0.1
1
2
3
4
5
6
7
8
9
10
MILES

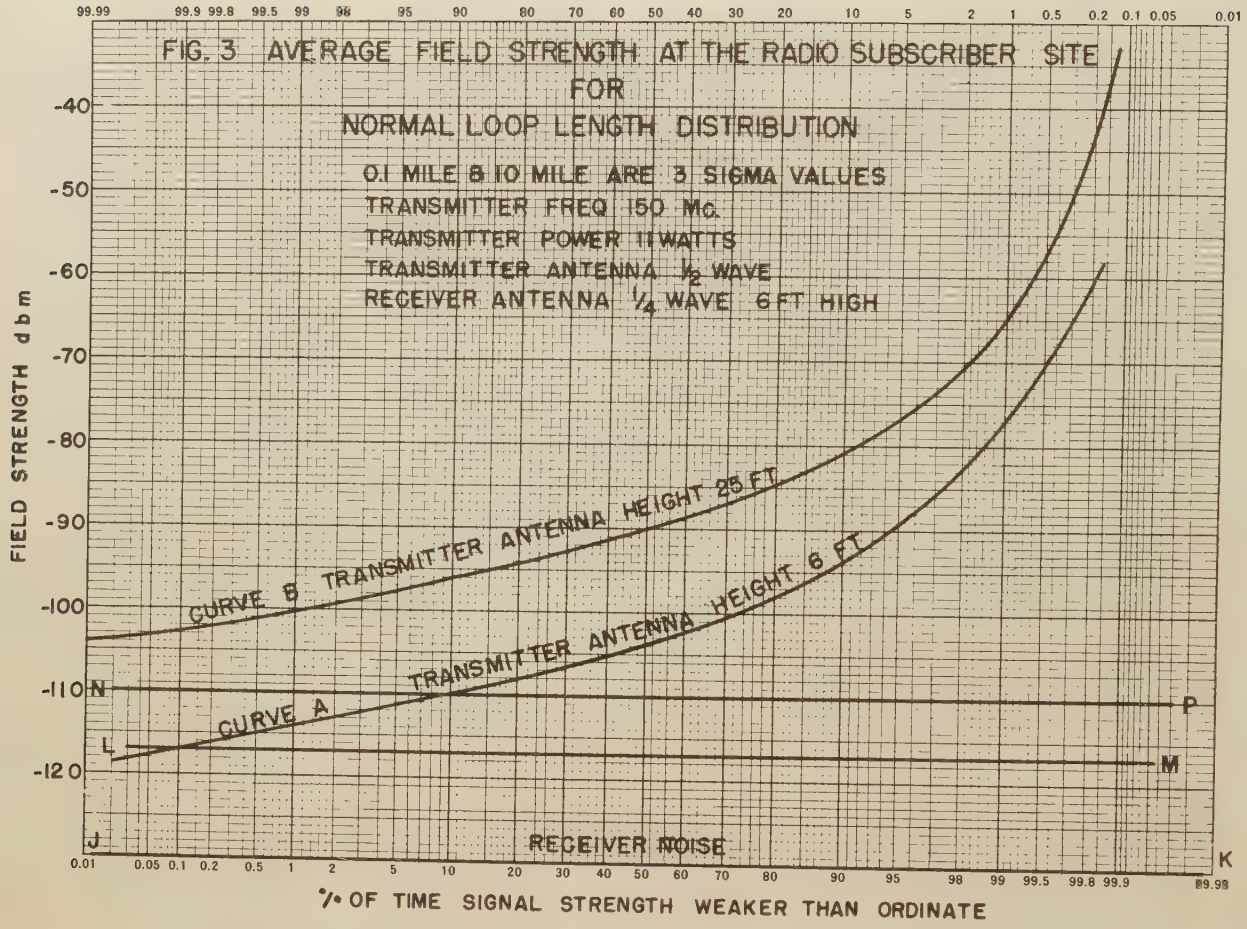
FIG. 3 AVERAGE FIELD STRENGTH AT THE RADIO SUBSCRIBER SITE FOR NORMAL LOOP LENGTH DISTRIBUTION
0.1 MILE & 10 MILE ARE 3 SIGMA VALUES
TRANSMITTER FREQ 150 Mc.
TRANSMITTER POWER 11 WATTS
TRANSMITTER ANTENNA ½ WAVE
RECEIVER ANTENNA ¼ WAVE 6 FT HIGH
CURVE B TRANSMITTER ANTENNA HEIGHT 25 FT.
CURVE A TRANSMITTER ANTENNA HEIGHT 6 FT.
RECEIVER NOISE
FIELD STRENGTH d b m
% OF TIME SIGNAL STRENGTH WEAKER THAN ORDINATE
99.99 99.9 99.8 99.5 99 98 95 90 80 70 60 50 40 30 20 10 5 2 1 0.5 0.2 0.1 0.05 0.01
0.01 0.05 0.1 0.2 0.5 1 2 5 10 20 30 40 50 60 70 80 90 95 98 99 99.5 99.8 99.9 99.99
-40 -50 -60 -70 -80 -90 -100 -110 -120
J K L M N P

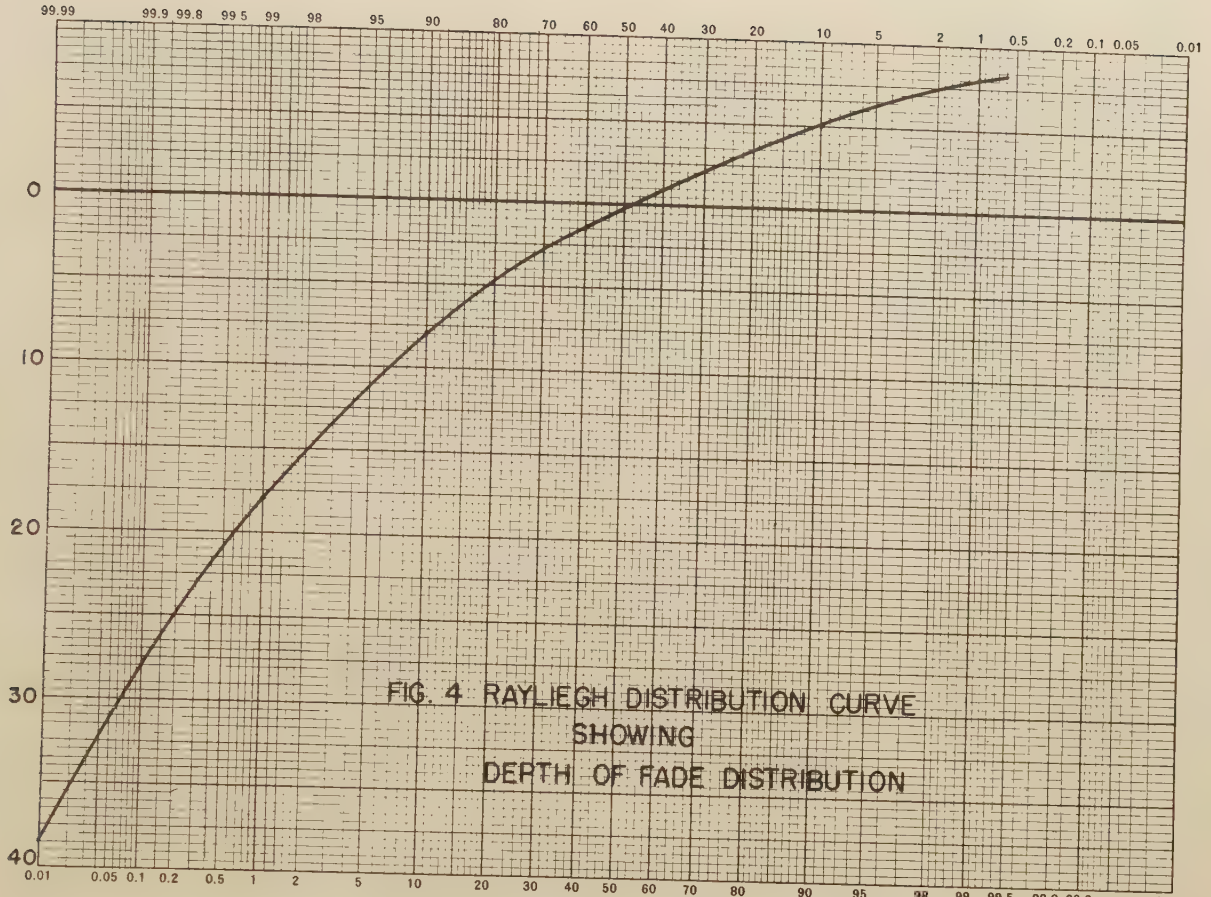

FIG. 4 RAYLIEGH DISTRIBUTION CURVE
SHOWING
DEPTH OF FADE DISTRIBUTION

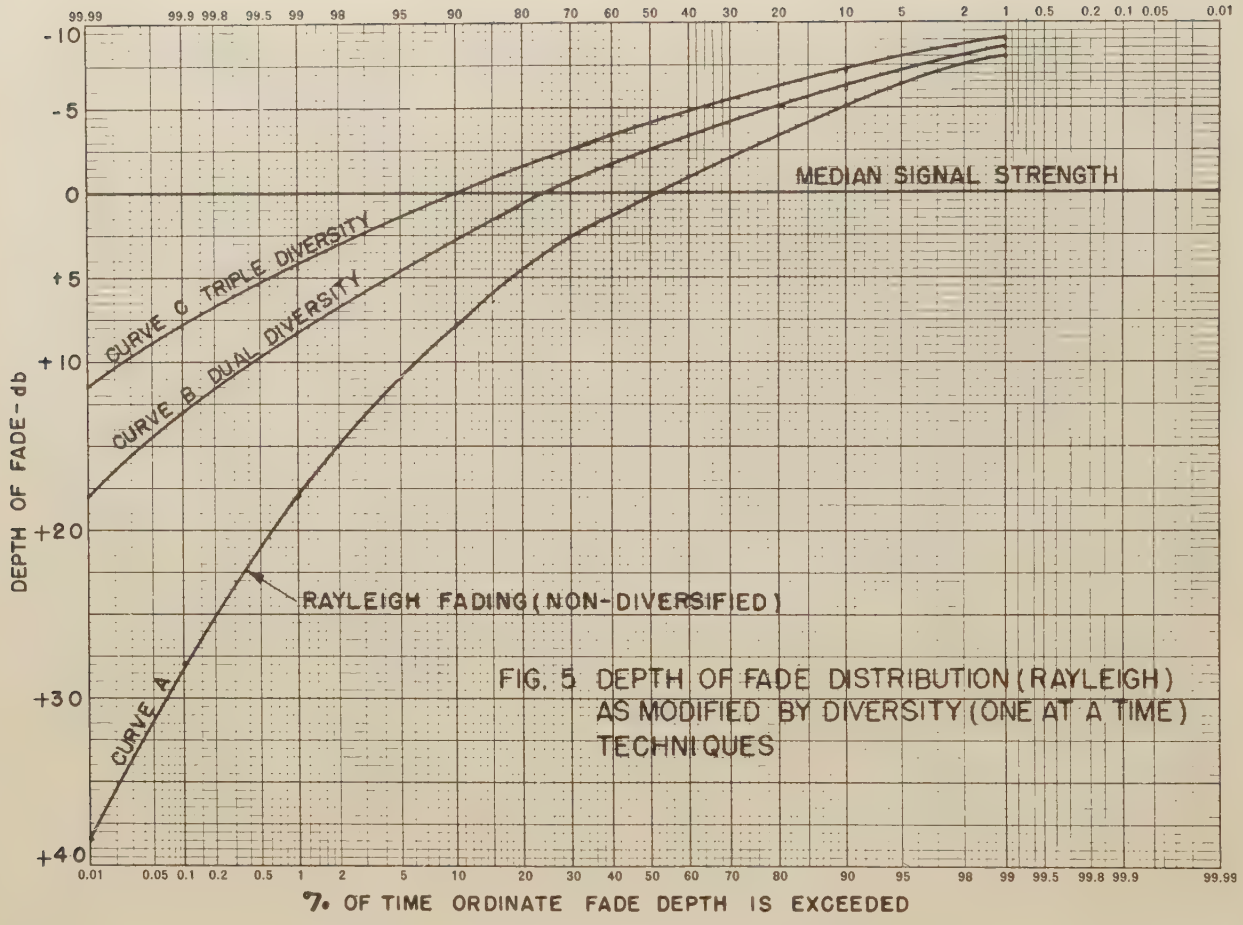
99.99 99.9 99.8 99.5 99 98 95 90 80 70 60 50 40 30 20 10 5 2 1 0.5 0.2 0.1 0.05 0.01
-10
-5
0
+5
+10
+20
+30
+40
DEPTH OF FADE - db
MEDIAN SIGNAL STRENGTH
CURVE C TRIPLE DIVERSITY
CURVE B DUAL DIVERSITY
RAYLEIGH FADING (NON-DIVERSIFIED)
CURVE A
0.01 0.05 0.1 0.2 0.5 1 2 5 10 20 30 40 50 60 70 80 90 95 98 99 99.5 99.8 99.9 99.99
% OF TIME ORDINATE FADE DEPTH IS EXCEEDED

FIG. 5 DEPTH OF FADE DISTRIBUTION (RAYLEIGH) AS MODIFIED BY DIVERSITY (ONE AT A TIME) TECHNIQUES

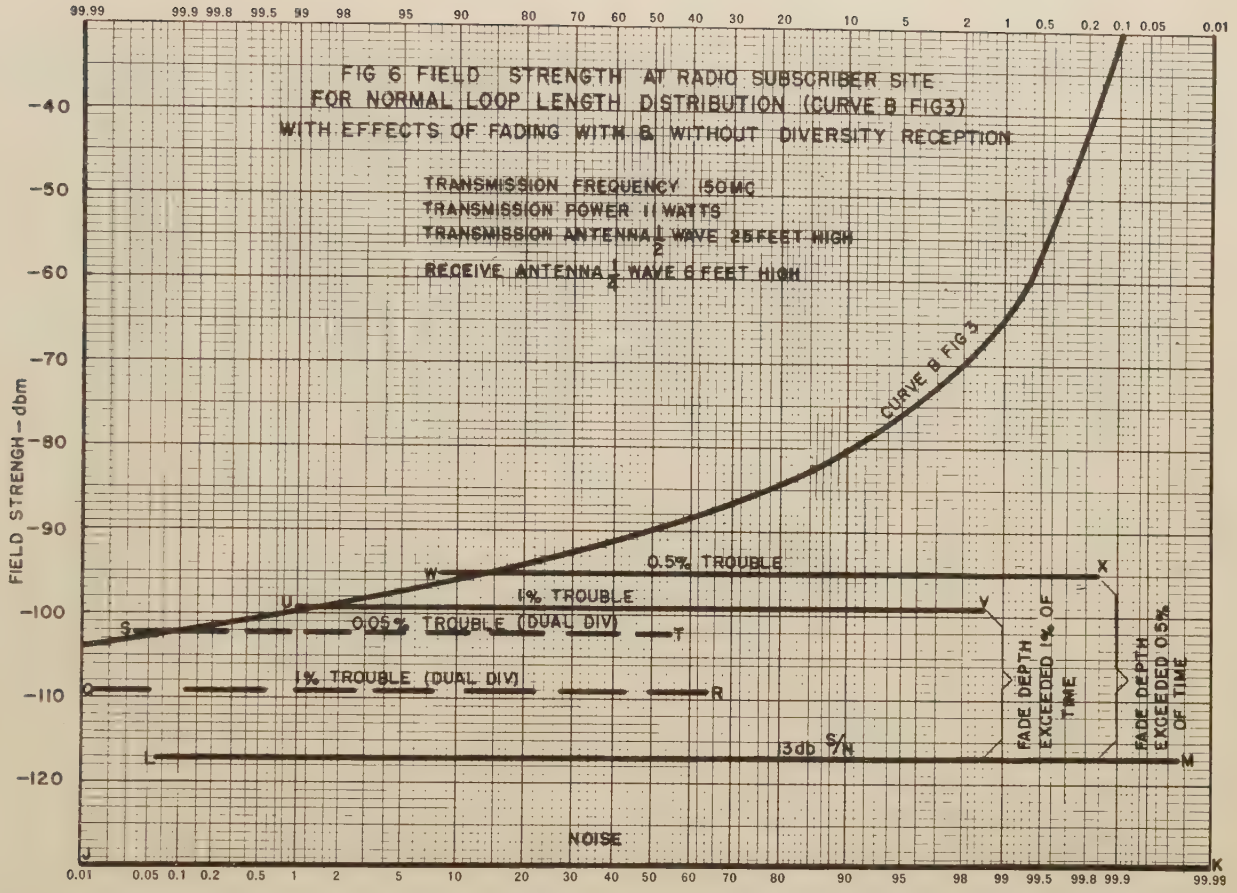
FIG 6 FIELD STRENGTH AT RADIO SUBSCRIBER SITE
FOR NORMAL LOOP LENGTH DISTRIBUTION (CURVE B FIG3)
WITH EFFECTS OF FADING WITH & WITHOUT DIVERSITY RECEPTION
TRANSMISSION FREQUENCY 150MC
TRANSMISSION POWER 11 WATTS
TRANSMISSION ANTENNA 1/2 WAVE 25 FEET HIGH
RECEIVE ANTENNA 1/4 WAVE 6 FEET HIGH
FIELD STRENGH - dbm
-40
-50
-60
-70
-80
-90
-100
-110
-120
99.99 99.9 99.8 99.5 99 98 95 90 80 70 60 50 40 30 20 10 5 2 1 0.5 0.2 0.1 0.05 0.01
0.01 0.05 0.1 0.2 0.5 1 2 5 10 20 30 40 50 60 70 80 90 95 98 99 99.5 99.8 99.9 99.99
CURVE B FIG 3
0.5% TROUBLE
1% TROUBLE
0.05% TROUBLE (DUAL DIV)
1% TROUBLE (DUAL DIV)
3db S/N
NOISE
FADE DEPTH EXCEEDED 1% OF TIME
FADE DEPTH EXCEEDED 0.5% OF TIME
W
X
U
V
S
T
Q
R
L
M
J
K

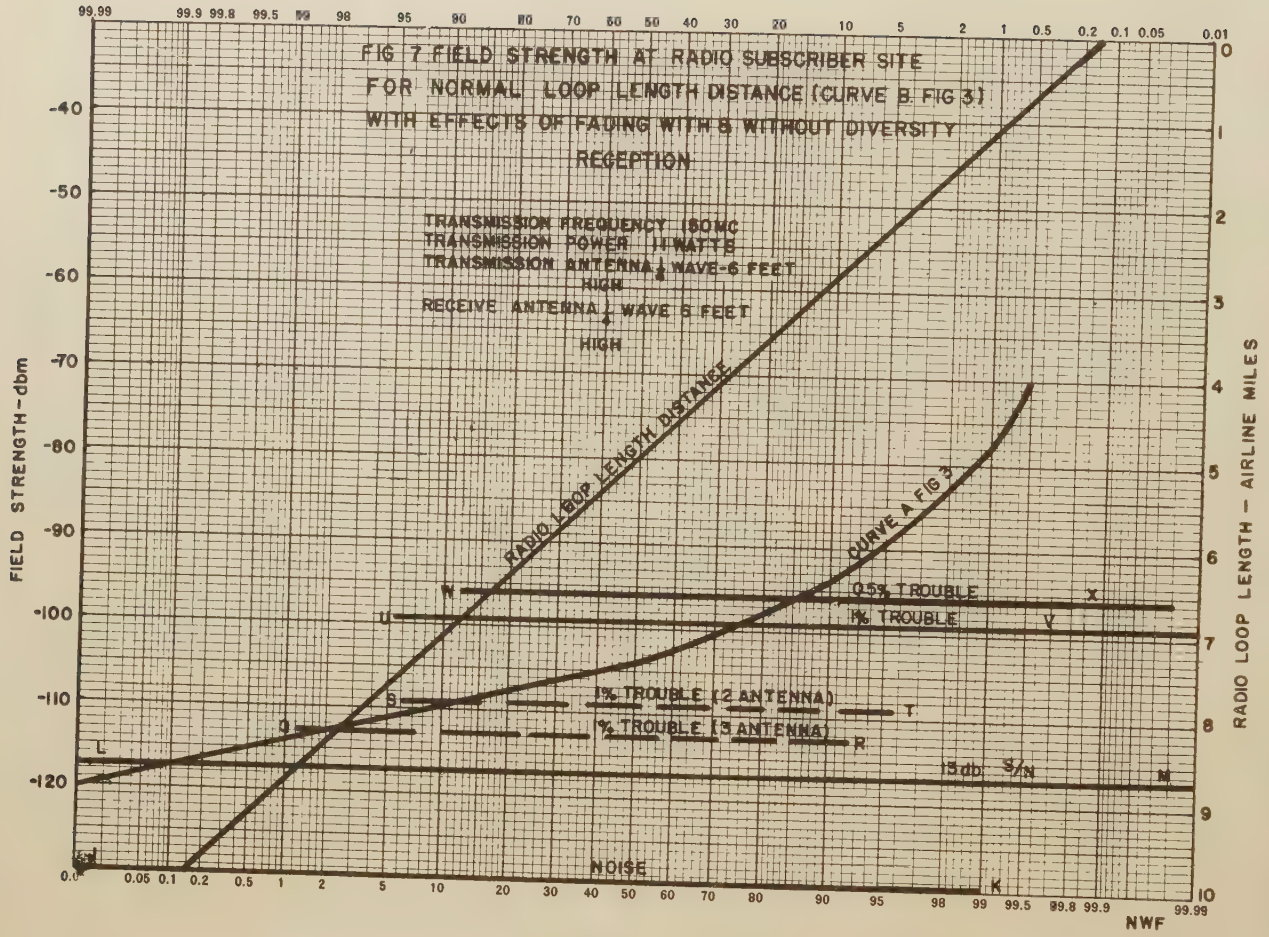
FIG 7 FIELD STRENGTH AT RADIO SUBSCRIBER SITE
FOR NORMAL LOOP LENGTH DISTANCE (CURVE B FIG 3)
WITH EFFECTS OF FADING WITH & WITHOUT DIVERSITY
RECEPTION
TRANSMISSION FREQUENCY 150MC
TRANSMISSION POWER 1 WATT
TRANSMISSION ANTENNA ¼ WAVE-6 FEET HIGH
RECEIVE ANTENNA ¼ WAVE 5 FEET HIGH
FIELD STRENGTH-dbm
RADIO LOOP LENGTH - AIRLINE MILES
RADIO LOOP LENGTH DISTANCE
CURVE A FIG 3
0.5% TROUBLE
1% TROUBLE
1% TROUBLE (2 ANTENNA)
1% TROUBLE (3 ANTENNA)
15db S/N
NOISE
NWF

FM INTERFERENCE AND NOISE-SUPPRESSION PROPERTIES OF THE OSCILLATING LIMITER

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Summary: An amplitude limiter with regenerative feedback can, under appropriate conditions, provide an effective tool for suppressing interference from undesired signals, as well as interstation noise. The amount of positive narrow-band feedback that is prescribed for marked improvement in the stronger-signal capture performance causes the limiter to oscillate in the absence of an input signal. Although oscillation results in automatic noise squelch, it also imposes limitations on the frequencies and minimum amplitudes of receivable signals, in the form of a locking frequency range and a locking threshold. The alleviation of these limitations requires careful design of the feedback phase characteristic. The theoretical discussion is followed by a summary of experimental data which brings out several important aspects of oscillating-limiter operation.

I. Introduction

Of the more serious problems that have plagued the users of FM systems, particularly in mobile communication services, interstation random noise and interference caused by multiple transmission paths and by other sources of spurious signals, consistently receive "the lion's share" of the receiver designer's attention. Noise and interference can be effectively eliminated if cost is no object. The design considerations that enter into interference rejection, however, have appeared to be entirely divorced from those for squelching interstation random noise. Consequently, separate efforts and expenditures have been made, even in the same design, to achieve receivers whose performance is relatively immune to both types of trouble. The necessity for establishing separate corrective circuits for each problem increases the cost of FM receivers considerably.

The two types of trouble can be alleviated by one simple scheme: a limiter with regenerative feedback. The basic theory of the effect of feedback around the limiter upon interference in FM reception has already been presented.[1] The present discussion is concerned with some important aspects of this technique that have not been covered in published work.

Two possible modes of operation can be distinguished for a limiter with regenerative band-limited feedback: oscillation-free operation and "oscillating-limiter" operation. Each mode of operation leads to interesting results.

The basic feedback theory[1] ignored the possible existence of a self-oscillation under regenerative feedback conditions. This will be justified in the present paper. A theory is presented here for studying the phenomenon of locking in oscillators that are built around an amplitude limiter when one or more signals are impressed at the oscillator input. Limitations imposed by phase shifts around the loop upon locking range and locking threshold will be discussed. The bearing of the results upon interference suppression in FM reception is emphasized. The discussion will include a resumé of experimental results.

II. Preliminary Considerations

The present discussion is concerned with the basic arrangement shown in Fig. 1. The idealized amplitude limiter is a device with associated frequency selectivity which transforms a sinusoid of amplitude E_{in}, whose frequency lies in the pass-band, into a sinusoid of the same frequency, whose amplitude E_{out} is described in Fig. 2(a) as a function of E_{in}. The ratio $E_{out}/E_{in} \equiv G_{\ell}$ is a voltage "amplification" which varies with E_{in}, as shown in Fig. 2(b). The amplitude-limiting

threshold is denoted by E_{th}. Clearly, for $E_{in} < E_{th}$, the idealized limiter performs as a linear amplifier with a voltage amplification of $G_{\ell\ max} = k/E_{th}$.

The phase shift around the loop will generally be a function of frequency. At some frequency, say the center of the passband, the signal that is fed back will arrive in phase with the input signal. Under restrictions (that will be discussed presently) on the amount of phase variation that can be tolerated over the desired passband the feedback will be termed regenerative over the passband if a zero-phase condition exists at some frequency in the passband (preferably the center frequency).

Thus, at the frequency of zero feedback phase shift, the loop gain is given by the product, $G_{fb} \cdot G_{\ell}$, of the transfer voltage ratio, G_{fb}, of the feedback network, and the transfer voltage ratio of the limiter itself. With reference to Fig. 1, when the switch S_{if} is closed and S_{fb} is open, the circuit will build up and sustain a stable oscillation if the feedback-loop gain $G_{loop} = G_{fb} \cdot G_{\ell} \geq 1$. Since $G_{\ell} = k/E_{in}$ for $E_{in} \geq E_{th}$, the voltage-transfer ratio of the limiter will vary with E_{in} when $E_{in} \geq E_{th}$, and it will subside to the value $G_{\ell o} = 1/G_{fb}$ when the condition of stable oscillation is reached. The frequency of the oscillation will clearly be the center frequency of the band, since the phase condition, by assumption, is satisfied at this frequency, and E_{in} will then be given by

$$E_{in} = E_{osc} = k\, G_{fb}. \tag{1}$$

For stability, and hence no self-sustained oscillation, the feedback voltage-transfer ratio at the frequency of exact in-phase feedback must satisfy the condition

$$G_{fb} < E_{th}/k \tag{2}$$

which is imposed by the maximum value that the limiter voltage-transfer ratio can attain. Accordingly, the amplitude $k\, G_{fb}$ of the voltage that is fed back must be smaller than the value E_{th} that is necessary to saturate the limiter. In view of this condition, the ideal limiters must in general be assumed to have a nonzero threshold of limiting, E_{th}, in order to allow for the possible existence of a condition of no self-sustained oscillation with nonzero values of G_{fb}.

In the ensuing discussion, we shall be concerned mainly with the properties of the resultant signal at the output of the system in Fig. 1 when the i-f signal is made up of two sinusoids. One of the sinusoids will be assumed to be $E_s \cos pt$, the other $a E_s \cos (p+r)t$, where $a < 1$ and $r \ll p$. The i-f bandwidth is assumed to be well defined by $(BW)_{if}$, and both p and p+r fall within the i-f passband. Thus,

$$e_{if}(t) = E_s[\cos pt + a \cos (p+r)t]$$

$$= E_s(1 + 2a \cos rt + a^2)^{1/2} \cos [pt + \theta(t)] \tag{3}$$

where

$$\theta(t) = \tan^{-1} \frac{a \sin rt}{1 + a \cos rt}. \tag{4}$$

The weaker-to-stronger-signal amplitude ratio a will frequently be referred to as the "interference ratio."

If $e_{if}(t)$ is impressed at the input of a limiter of bandwidth $(BW)_{lim}$, then, provided that

$$(1-a)\, E_s \geq E_{th}, \tag{5}$$

the signal at the output of the limiter will be

$$e_{out}(t) = k \sum_{n=-N}^{M} A_n(a) \cos (p - nr)t \tag{6}$$

where $kA_n(a)$ is the amplitude of the spectral component at frequency p-nr rad/sec in the structure of the amplitude-limited resultant of the two input sinusoids. N represents the number of sideband components whose frequencies are greater than p, and M represents the number of lower sideband components that are passed by the limiter filter. The numbers N and M will depend upon the position of p within the i-f passband, and upon $(BW)_{lim}$ and the frequency difference r. Thus if $(BW)_{if} \geq r > (BW)_{if}/2$, and $(BW)_{lim} = (BW)_{if}$, $M = 0$ and $N = 1$. Under these conditions, the amplitude-limited resultant of the two sinusoids will be said to experience the condition of most

severe frequency-band limitation.

If the phase shift around the loop in Fig. 1 can be ignored, it has been shown[1] that when the feedback-loop bandwidth equals $(BW)_{if}$ and r lies in the range $(BW)_{if} \geq r > (BW)_{if}/2$, positive feedback reduces the interference ratio from the value $\underline{a}$ at the output of the i-f amplifier to the value a_ℓ at the input of the limiter. The amount of reduction in the interference ratio is then given by

$$\Delta a \equiv a - a_\ell = K_s f_1(a_\ell), \tag{7}$$

where

$$f_1(a_\ell) = a_\ell A_o(a_\ell) - A_{-1}(a_\ell)$$

and

$$K_s = k G_{fb}/E_s. \tag{8}$$

The function $f_1(a_\ell)$ is plotted in Fig. 3. Figure 3 also presents a plot of $f_2(a_\ell)$ which applies when the feedback is taken around two cascaded limiters, the first limiter having a bandwidth of $(BW)_{if}$, but otherwise under the same feedback and interference conditions as for $f_1(a_\ell)$.

Comparison of $f_1(a_\ell)$ and $f_2(a_\ell)$ reveals that a greater reduction in the interference ratio is achieved between the output of the i-f amplifier and the input of the first limiter, when the regenerative band-limited feedback is taken around two narrow-band limiters in cascade rather than around one limiter. Similar reasoning reveals the validity of the following statements[2, 3]:

Theorem I. If the feedback is taken from the output of the second, third, or n^{th} stage of ideal limiting to the input of the first (each stage, except possibly the $(n-1)^{st}$, having one i-f bandwidth) and the feedback bandwidth is equal in extent to one i-f bandwidth, then, under the most adverse interference conditions delivered by the i-f amplifier, the effect upon the capture performance, caused by the increase in the amplitude difference between the two signals with positive feedback and the decrease in amplitude difference with inverse feedback, becomes increasingly more pronounced with higher values of n.

Corollary. The longer the chain of narrow-band limiters that is spanned by the feedback branch, the more insignificant the amplitude of the weaker signal relative to the stronger signal becomes with positive feedback; and the closer together in amplitude the two signals will be with inverse feedback, both signals being observed at the input of the first limiter in the chain.

Theorem II. When the feedback is taken through a feedback bandwidth of one i-f bandwidth, from the output of the n^{th} limiter in a chain to the input of the first limiter, the effect of the feedback upon the ratio of weaker-to-stronger-signal amplitude at the input of the first limiter is greatest when each limiter, up to and including the $(n-1)^{st}$, possesses its lowest permissible bandwidth value of one i-f bandwidth.

Theorem III. When the feedback bandwidth is the only parameter subject to variation, the effect of the feedback is greatest when the feedback bandwidth takes on its lowest permissible value of one i-f bandwidth.

The application of regenerative feedback around the limiter also relaxes the condition on the signal level that is necessary for limiter saturation. Thus if only one sinusoid of amplitude E_s is delivered by the i-f amplifier, then in the absence of feedback around the limiter (i.e., with S_{fb} closed) the condition is $E_s \geq E_{th}$. If, however, E_s is less than E_{th}, and if the feedback phase shift at the frequency of the signal is zero, then at the instant of opening S_{fb}, $E_{fb} = k E_s G_{fb}/E_{th}$, and we demand that

$$E_s \left[1 + \frac{k G_{fb}}{E_{th}}\right] \geq E_{th}. \tag{9}$$

Thus, in the presence of regenerative feedback, the threshold of limiting, as seen from the output of the i-f amplifier, is decreased from E_{th} to $E_{th}/(1 + k G_{fb}/E_{th})$. The effect of nonzero phase shift around the loop is to increase the threshold beyond this new value. It is also interesting to note that once the limiter operating point has been brought into the saturation region, the condition

on the input signal amplitude, in order to keep the limiter saturated, becomes

$$E_s(1 + K_s) \geq E_{th}, \tag{10}$$

which represents a lower threshold than the one indicated in condition (9). The parameter K_s is defined in Eq. (8).

When the i-f signal is given by Eq. (3), the limiting-threshold condition in the absence of feedback is given by (5). When regenerative feedback is applied, the placement of the limiter operating point in the saturation region requires that

$$E_s \geq \frac{E_{th}}{(1-a)(1 + k G_{fb}/E_{th})} . \tag{11}$$

Once the proper operation has been established, the signal level can dip below the threshold indicated in (11), but it must satisfy the new condition

$$E_s \geq \frac{E_{th}}{(1-a) + K_s[A_o(a_\ell) - A_{-1}(a_\ell)]}, \tag{12}$$

in which we have ignored the effect of phase shift off the center frequency, and assumed that the feedback bandwidth equals $(BW)_{if}$.

III. Oscillation-Free Operation

In order to ensure oscillation-free operation when regenerative feedback is applied around the limiter, the feedback voltage-transfer ratio at the frequency of zero feedback phase shift must satisfy the condition $k G_{fb}/E_{th} < 1$. This condition can be rewritten in the form

$$K_s \equiv k G_{fb}/E_s < E_{th}/E_s. \tag{13}$$

An upper bound on E_{th}/E_s can be obtained from the condition that is necessary for the placement of the limiter operating point in the saturation region. Thus, in the absence of interference, condition (9) leads to

$$E_{th}/E_s \leq \left[1 + (1 + 4K_s)^{1/2}\right]/2$$

which, in conjunction with (13), yields

$$K_s < 1 + \sqrt{3/2} = 2.225. \tag{14}$$

In order to understand the significance of this condition, let us note from Eq. (8) that K_s is the ratio of the amplitude, E_{fb}, of the signal that is fed back to the amplitude, E_s, of the signal delivered by the i-f amplifier. Since E_s can range from zero to some very large value, the value of K_s will generally lie between ∞ and some very small value. But if E_s is too low (which corresponds to very large K_s), condition (9) may not be satisfied. Consequently, the limiter would not limit and the system would operate as a stable linear amplifier with regenerative feedback. Thus the restriction indicated in (14) simply means that the system will operate as a limiter with regenerative feedback for all input signal amplitudes E_s that exceed $0.45\, E_{fb}$.

After the establishment of the condition for proper operation, condition (10) applies, and the value of K_s becomes bounded by the condition

$$K_s < \frac{E_{fb}}{E_{th} - E_{fb}},$$

with $E_{fb} = k G_{fb}$.

In the presence of interference, let the i-f signal be described as in Eq. (3). Condition (11) applies, and it leads to

$$\frac{E_{th}}{E_s} < \frac{1}{2}(1-a)\left\{1 + [(1 + K_s/(1-a)]^{1/2}\right\}.$$

This condition, in conjunction with (13), leads to

$$K_s < (1-a)/0.45. \tag{15}$$

Once a feedback steady state has been established, condition (12) applies under the conditions specified with it, and K_s becomes bounded by

$$K_s < \frac{1-a}{1 + A_{-1}(a_\ell) - A_o(a_\ell)} . \tag{16}$$

A similar analysis can be carried out for regenerative feedback around two or more limiters. Although the results indicate tighter restrictions on K_s, the stronger-signal enhancement effect that results from regenerative feedback around two or more limiters grows sufficiently more pronounced to offset the increased restrictions.

It should be clear from the preceding results

that the requirement of stable, oscillation-free operation imposes severe restrictions on the value of K_s because it brings in the question of limiter saturation. These restrictions on K_s signify severe limitations on the weaker-signal suppression performance of a limiter with regenerative band-limited feedback, since this performance improves markedly with increased K_s.

IV. The Oscillating Limiter

Let us now consider the situation in which the feedback-loop gain is sufficiently high for the establishment and maintenance of oscillation. Since in this condition $E_{fb} = k G_{fb} \geq E_{th}$, questions of limiter saturation do not arise. As we have mentioned, the frequency of the self-oscillation will be the frequency of exact in-phase feedback. The amplitude of the steady oscillation at the limiter input will be $k G_{fb}$.

With reference to Fig. 1, it is interesting to note that two approaches can be taken in the study of the system when $k G_{fb} \geq E_{th}$. First, we can start with S_{if} open and S_{fb} closed, and proceed to determine what happens after we open S_{fb}. No self-oscillation will be present before and just after S_{fb} is opened. The situation is therefore handled purely as a feedback problem in which no signals other than the excitation described by $e_{if}(t)$ are present. [1]

An equally interesting approach, which in a sense complements the feedback argument, starts with S_{fb} open and S_{if} closed. Thus, after a steady oscillation has been established, the switch S_{if} is opened to introduce an external signal source. This signal source is assumed to be an ideal voltage source.

The introduction of an external signal will change the amplitude of the resultant signal at the input of the limiter. This will shift the operating point on the output-input amplitude characteristic of the limiter (Fig. 2a), and thereby alter the value of G_ℓ. If the input signal is a sinusoid of amplitude E_s whose frequency equals the frequency of the self-oscillation, then, regardless of the initial phase relationship between the signal and the oscillation, the phase of the resultant signal at the input of the limiter will always settle down to the phase of the impressed signal, and its amplitude will be $E_s + E_{osc} = E_s[1 + K_s]$, where $K_s = k G_{fb}/E_s$. The result of increasing E_{in} beyond the self-oscillation amplitude E_{osc} is to decrease the amplification of the limiter below the value necessary for an over-all feedback-loop gain of unity. Consequently, the presence of the impressed signal causes the system to operate as a nonoscillating limiter with regenerative feedback. These same conclusions apply if the impressed sinusoid is allowed to carry a slow frequency modulation, $\Delta\Omega g(t)$, where $|g(t)|_{max}=1$, as long as the feedback phase shift deviates negligibly from the in-phase condition over the frequency range covered by the signal, and $\Delta\Omega$ is a small fraction of the feedback bandwidth.

Now let the frequency-modulated sinusoid be deviated over the entire i-f bandwidth. Assume that S_{if} is opened at the instant when the signal frequency is momentarily equal to $p = \omega_o - r$ rad/sec, where ω_o is the oscillation frequency (the center of the band) and $r \ll p$. Let $r > (BW)_{if}/4$. The feedback bandwidth is assumed hereafter to be sharply defined and equal to $(BW)_{if}$. Denote by $K_s E_s$ the amplitude of the self-oscillation, before and at the instant of switching. As before, $K_s = k G_{fb}/E_s$. Two situations will be distinguished: the impressed signal is stronger than the oscillation ($K_s < 1$), and it is weaker ($K_s > 1$).

Consider, first, the situation in which $K_s < 1$. At the instant of opening the switch S_{if} in Fig. 1, the resultant signal at the input of the limiter is given by

$$E_s \cos pt + K_s E_s \cos (p+r)t.$$

In response, the feedback voltage becomes

$$k G_{fb}\{A_o(K_s) \cos (pt - \phi_{fb}) + A_{-1}(K_s) \cos (p+r)t$$

$$+A_{-2}(K_s) \cos [(p + 2r)t + \phi_{fb}]\},$$

where ϕ_{fb} is the phase shift around the loop at r rad/sec off the center frequency $\omega_o = p+r$. For simplicity, let us first assume that ϕ_{fb} is small, and that $K_s|A_{-2}(K_s)|$ is negligible compared with

$1 + K_s A_o(K_s)$. The second of these is a good assumption for all $K_s < 1$ because $K_s|A_{-2}(K_s)|$ is less than 1/10 of $1 + K_s A_o(K_s)$ for all $K_s \leq 0.96$, and the ratio goes to $1/(8.45)$ as K_s goes to unity.

Thus, at the beginning of the second round (neglecting ϕ_{fb} for the time being), we have a resultant voltage at the input of the limiter whose components have amplitudes of

$$E_s + k G_{fb} A_o(K_s) \quad \text{at } p$$

and

$$k G_{fb} A_{-1}(K_s) \quad \text{at } \omega_o = p+r$$

Compared to the amplitude of the component at the frequency of the impressed signal, the amplitude of the component at the oscillation frequency is now given by

$$K_s \frac{A_{-1}(K_s)}{1 + K_s A_o(K_s)} = K_s f(K_s).$$

As K_s approaches unity, $A_{-1}(K_s)$ approaches $2/\pi$ from below, and $A_o(K_s)$ approaches $2/\pi$ from above. Therefore, the maximum value of $f(K_s)$, which occurs at $K_s = 1$, is approximately 0.39. Thus, the ratio of the self-oscillation amplitude to the amplitude of the impressed signal undergoes a sizable reduction, in the first round, from K_s to something less than $0.4K_s$. If the ratio at the beginning of the second round is denoted by K_2, then at the beginning of the third round K_2 is reduced by a factor of $f(K_2)$, which is considerably less than unity.

As the signal flow is traced around the loop, it is immediately realized that the amplitude of the self-oscillation quickly goes to zero while the amplitude of the impressed signal quickly builds up to $E_s(1 + K_s)$, the value that may be predicted by reasoning purely in terms of feedback, ignoring the self-oscillation. This quick decay of the self-oscillation and the accompanying transfer of the energy, initially residing in it, to the frequency of the impressed forcing excitation, is a manifestation of a "locking" phenomenon. The introduction of the signal at p, whose amplitude was assumed greater than the amplitude of the self-oscillation, will quickly pull the oscillation from the frequency p+r to the frequency p, and literally superimpose the amplitude of the oscillation upon the amplitude of the impressed signal at the frequency p. In this situation, the decay of the self-oscillation and the shift of its energy to the frequency of the impressed signal can be ascribed to the fact that the amplification of the limiter is decreased below the value necessary for the maintenance of a self-oscillation around the loop.

If the feedback angle ϕ_{fb} at the frequency p of the impressed signal cannot be neglected, then at the end of the first round around the loop, the feedback component having the frequency p of the impressed signal will have an amplitude given by $k G_{fb} A_o(K_s)$ but its phase will deviate from that of the impressed signal by ϕ_{fb}. The component at the frequency of the self-oscillation will of course sustain no phase shift around the loop. The ratio of amplitudes at the beginning of the second round, therefore, will be

$$K_s \frac{A_{-1}(K_s)}{\left[1 + 2K_s A_o(K_s) \cos \phi_{fb} + K_s^2 A_o^2(K_s)\right]^{1/2}} = K_s f'(K_s)$$

It is easy to compute the range of ϕ_{fb} in which $f'(K_s)$ will be less than unity. But this is not necessary, since $A_{-1}(K_s) < 1$ for all $K_s \leq 1$, and the denominator will exceed unity for all $\cos \phi_{fb} > -(1/2) K_s A_o(K_s)$. Since the advantages of positive feedback relate fundamentally to the fact that it increases the amplitudes at the input rather than subtracts from them, it is logical to desire that the angle ϕ_{fb} be smaller in magnitude than $\pi/2$ throughout that range of the feedback passband which will be occupied by the expected signal carrier. For values of ϕ_{fb} satisfying these restrictions, the maximum value of $f'(K_s)$ will always be bounded by 0.54, which is the value for $K_s = 1$ and $\phi_{fb} = \pi/2$. This again indicates a considerable decrease in the amplitude of the self-oscillation in the first round. A steady state will, therefore, be reached quickly in which the energy in the oscillation frequency is shifted to the frequency of the impressed signal. The pertinent

phase relationships in this steady state are shown in Fig. 4. The phasor OR represents the resultant amplitude of the source-frequency signal appearing across the input terminals of the limiter. The energy originally oscillating at the zero phase-shift frequency of the feedback passband, now resides in a sinusoid at the frequency p of the impressed signal and shifted in phase relative to this signal by $\phi_{fb} + \psi$. The angle ψ is given by $\sin \psi = K_s \sin \phi_{fb}$, and hence ψ is a real angle for all

$$K_s \sin \phi_{fb} \leq 1 \qquad (17)$$

which is ensured by our initial assumption that $K_s < 1$.

Note that at the frequency of the oscillation, $\phi_{fb} = 0$ and hence $\psi = 0$. This shows that regardless of the initial phase relationship between an impressed sinusoid whose frequency equals ω_o and the self-oscillation, the resultant signal at the limiter input will eventually be exactly in phase with the impressed sinusoid, and its amplitude will be $E_s[1 + K_s]$.

Let us now consider the situation in which the amplitude of the impressed signal at p is smaller than the amplitude of the self-oscillation at $\omega_o = p+r$. Thus, if the amplitude of the impressed signal is given by E_s, let the amplitude of the self-oscillation be given by

$$k G_{fb} = E_s/\alpha$$

where $\alpha < 1$. Again, we start by assuming that the angle ϕ_{fb} is negligible. At the end of the first round, after opening the switch S_{if} (in Fig. 1), the resultant signal at the input of the limiter is made up of the components

$E_s[1 + (1/\alpha) A_{-1}(\alpha)]$ at p rad/sec

$E_s A_o(\alpha)/\alpha$ at p+r rad/sec

$E_s A_1(\alpha)/\alpha$ at p + 2r rad/sec.

The amplitude-limiting operation thus gives rise to a new component at p + 2r rad/sec. The ratio of the amplitude of the component at p + 2r to the amplitude of the component at p has a maximum value of 1/3 for $\alpha = 0$, and decreases monotonically to 0.13 as $\alpha \to 1$. The amplitude ratio of the signal component at p to the self-oscillation component at p+r increases from zero for $\alpha = 0$, to 1 for $\alpha = 0.593$, to 2.57 for $\alpha = 1$, and it always exceeds $(3/2)\alpha$. This indicates a significant boost in the amplitude of the component at the signal frequency at the expense of the self-oscillation.

In the second round around the feedback loop, the resultant of the three sinusoids is first amplitude-limited. Again, only three components from the spectrum of the amplitude-limited resultant signal are fed back. At the end of the second round, the amplitudes of the components of the resultant signal at the input of the limiter become

$E_s[1 + (1/\alpha) Q_{-1}(\alpha)]$ at p rad/sec

$E_s Q_o(\alpha)/\alpha$ at p+r rad/sec

$E_s Q_1(\alpha)/\alpha$ at p + 2r rad/sec,

where the Q's are computed from a Fourier analysis of the amplitude-limited resultant of three sinusoids of specified relative amplitudes and relative phases. The amplitude of the component at the signal frequency will again be increased at the expense of the self-oscillation component. If additional rounds are traced around the loop, the component at the signal frequency eventually emerges predominant over the oscillation component. Once this happens, the oscillation component and the spurious component at p + 2r deteriorate quickly and the balance of the energy that was originally in the oscillation shifts to the frequency of the signal.

Let us illustrate with the specific situation in which $\alpha = 0.4$. The results of tracing the first three rounds after the opening of S_{if} are summarized in Table I. It is clear that in the present instance the signal component predominates after the third round. In other instances, as $\alpha \to 0$, more rounds must be traced before the component at the applied signal frequency overtakes the oscillation and avalanches to complete control. But when $\alpha > 0.593$, the component at the signal frequency will dominate the oscillation component at the end of the first round.

TABLE I

Amplitudes of Components of Resultant Signal at the Input of the Limiter When $a = 0.4$

Frequency of Component	Amplitudes (Relative to Strongest Component) at the End of		
	First Round	Second Round	Third Round
p	0.630	0.754	1.000
p+r	1.000	1.000	0.926
p + 2r	0.196	0.216	0.173

Thus, the energy in the self-oscillation is quickly pulled over to the frequency of the impressed signal, even when the impressed signal is initially much smaller in amplitude than the self-oscillation. Incidentally, at the end of the pulling process, the resultant amplitude of the signal at the impressed frequency is given by

$$E_s + k G_{fb} = E_s(1 + K_s), \quad K_s > 1,$$

the exact value that could have been written from reasoning on the feedback basis which ignored the initial existence of the self-oscillation.

If the preceding analysis is repeated with the spurious component at p + 2r ignored (as would be justifiable if the oscillation frequency did not fall at the center of the band and $r > (BW)_{if}/2$), the same conclusions would have been reached with much less computational labor. Thus, with the p + 2r component rejected, the ratio of the amplitude of the component at the frequency p of the source to the amplitude of the component of self-oscillation at p+r, as seen by the limiter, becomes at the end of the n^{th} round

$$a_{n+1} = \frac{a_n + A_{-1}(a_n)}{A_o(a_n)} .$$

Note that a_{n+1} always exceeds $(3/2)a_n$ when $a_n < 1$. The value of a_n for which a_{n+1} equals unity is 0.593. Beyond this value of a_n the amplitude of the signal overtakes the amplitude of the self-oscillation. For if $a_{n+1} = 1$, $a_{n+2} = 2.57$. Evidently, the effect of including the component at p + 2r (when it is passed) in the computation is to slow down the convergence of the process of transfer of the oscillation energy to the frequency of the impressed signal.

If the feedback phase shift ϕ_{fb} at the frequency of the impressed signal is taken into account, the argument is basically unchanged. Thus, at the end of the first round after S_{if} is opened, the amplitude ratio of the component at the signal freguency to the component at the oscillation frequency is

$$a_1 = \frac{a}{A_o(a)} \left(1 + (2/a)\, A_{-1}(a) \cos \phi_{fb} + A_{-1}^2(a)/a^2\right)^{1/2}$$

$$\approx a\left(\frac{5}{4} + \cos \phi_{fb}\right)^{1/2} \qquad \text{for } 0 \leq a \leq 0.63.$$

The quantity in parentheses exceeds unity for all values of $\cos \phi_{fb} > -A_{-1}(a)/2a$. With this restriction on ϕ_{fb}, a_1 will always exceed a, which indicates a boost in the signal amplitude relative to the amplitude of the self-oscillation. As additional rounds are traced around the feedback loop, the resultant signal phasor, as seen at the limiter input, describes a pattern similar to that shown in Fig. 5. The component that is fed back at the signal frequency continues to grow in this diagram until a steady state is reached in which the resultant signal component is given by OR_∞. If the signal-to-oscillation amplitude ratio at the limiter input is given by a_n at the end of the n^{th} round, and if a_n is less than unity, the amplitude of the component that is fed back at the signal frequency in the $(n+1)^{th}$ round will be given by $k G_{fb} Q_{-1}(a_n) = E_s Q_{-1}(a_n)/a$, where $Q_{-1}(a_n)$ is the amplitude of the Fourier component at the frequency of the applied signal in the structure of

the amplitude-limited resultant of the three sinusoids (at p, p+r, and p + 2r rad/sec). If $a_n > 1$, Q_{-1} is replaced by $Q_o(1/a_n)$, and $Q_o(x) \to 1$ as $x \to 0$, which indicates that the amplitude of the fed back component approaches $k G_{fb}$, the full amplitude of the initial self-oscillation. For $a_n > 1$, the amplitude of the component at the self-oscillation frequency dies out according to the function $Q_{-1}(x)$ which goes to zero as $x \to 0$.

If a steady state in which the self-oscillation energy is properly shifted to the frequency of the impressed signal exists, the phase relationships indicated in Fig. 4 (or by triangle $O E_s R_\infty$ in Fig. 5) must be satisfied. Accordingly, we must demand that $K_s \sin \phi_{fb} \leq 1$. When $K_s > 1$, this condition restricts the permissible values of ϕ_{fb} for given values of K_s (thus imposing a locking frequency range), and it restricts K_s for given ϕ_{fb} (thus imposing a locking threshold). If this condition is not satisfied, the phase of the sinusoid that carries the self-oscillation energy will not be fixed relative to the impressed sinusoid, but will continue to slip. The method of analysis whereby the signal flow around the loop is traced around and around will never converge to a repeatable pattern. We may, therefore, conclude that no true locking exists if condition (17) on K_s and ϕ_{fb} is not satisfied.

V. Interpretation of the Locking Mechanism

The open-loop transfer function of the system shown in Fig. 1 is essentially the transfer function of a fixed-parameter filter, scaled by the magnitude factor $G_\ell = k/E_{in}$ which is contributed by the saturated limiter. In the absence of an applied signal, the system will oscillate at the frequency ω_o of exact in-phase feedback, with an amplitude that corresponds to $E_{in} = E_{osc} = k G_{fb}(\omega_o)$ if $k G_{fb}(\omega_o) \geq E_{th}$. With the assumed idealizations, $k|G_{fb}|$ is independent of frequency over a well-defined passband of one i-f bandwidth, and the frequency ω_o of exact in-phase feedback is fixed. The presence of an applied signal can only change the value of the magnitude factor G_ℓ which is contributed by the idealized limiter. An "oscillating limiter" will "lock" to an applied signal by ceasing to oscillate and by operating as a stable system with regenerative feedback if the presence of the applied signal causes the amplitude E_{in} of the resultant signal at the input of the limiter to increase beyond the value $E_{in} = E_{osc} = k G_{fb}$, which is necessary for a feedback-loop gain of unity.

In contrast, there are oscillating systems that will lock to an applied signal if this signal is able to change the frequency of exact in-phase feedback so that it coincides with the signal frequency. In such systems, the applied signal can be said to introduce, or change the value of, a simulated reactance that changes the frequency of the resonator. If the feedback-loop gain remains fixed (or is maintained at the value of unity) the oscillator can be said to lock to the applied signal by being forced to oscillate at the frequency of the applied signal.

It is also possible for the applied signal to reduce the feedback-loop gain, as well as to change the frequency of exact in-phase feedback of an oscillating system. Such a system belongs fundamentally to the class of oscillating limiters and, with appropriate design, it represents an oscillating limiter in which one possible measure has been taken to widen the locking range and lower the locking threshold.

Thus we are distinguishing here between two fundamentally different locking mechanisms. A "locked oscillating limiter" does not oscillate while it is locked to the frequency of an applied signal. But when this signal is removed it will oscillate. A locked oscillator oscillates all the time, but its frequency of oscillation is shifted by the controlling applied signal without destroying the gain condition for oscillation.

We have shown previously[1] that when the applied sinusoid has amplitude E_s and a frequency at which $\phi_{fb} \neq 0$, a feedback steady-state will exist only if condition (17) is satisfied. It is interesting to note that if we seek the condition for the amplitude of the resultant signal (at the input of the limiter in the feedback steady state) (see Fig. 4 or Fig. 5) to exceed the value

$E_{in} = k G_{fb} = E_{osc}$ which is necessary for unit feedback-loop gain, this condition turns out to be

$$2 K_s \sin (\phi_{fb}/2) < 1 . \tag{18}$$

This condition is evident from the constructions made in Fig. 6. If condition (17) is rewritten in the form

$$2 K_s \sin (\phi_{fb}/2) \cos (\phi_{fb}/2) < 1 , \tag{19}$$

or if the constructions in Fig. 6 are examined closely, it becomes obvious that the satisfaction of condition (18) guarantees the satisfaction of condition (17); but the reverse is not true. Since $\cos (\phi_{fb}/2) \geq 0.99$ for $|\phi_{fb}| \leq 16.33°$, and $\cos (\phi_{fb}/2) \geq 0.95$ for $|\phi_{fb}| \leq 36.33°$, the two conditions are substantially equivalent over a wider range of feedback phase shifts than should be encountered within the desired passband in a well-designed system.

Since condition (18) is more stringent than condition (17), we might inquire how condition (17) can rule out the existence of a self-sustained oscillation when, under circumstances in which it is satisfied, the limiter operates at a point which enables the feedback loop gain to exceed or equal unity. The answer to this is that a feedback steady state is really possible only after the self-oscillation has died out. With reference to Fig. 5, note that the resultant-signal phasor acquires amplitudes that exceed the steady-state amplitude $\overline{OR}_\infty$ before it settles down to OR_∞. The system operates in a condition of "stable equilibrium" once the steady-state condition has been reached. Any attempt by a self-oscillation to regenerate will raise the amplitude level at the limiter input, and thereby undermine the loop-gain condition that is necessary for the existence of a self-sustained oscillation.

VI. Locking in the Presence of Interference

Suppose a linear combination of two signals, as described in Eq. (1), is impressed across the terminals of the switch S_{if} (in Fig. 1). We shall consider in detail only the two extreme situations in which the frequency difference, r, is either a small fraction of the feedback bandwidth (which by assumption is well-defined by one $(BW)_{if}$) or a large fraction of it (i.e., $r > (BW)_{if}/2$).

If the frequency difference satisfies the condition[4]

$$r \leq r_{min} = \rho \frac{1-a}{1+a} (BW)_{if},$$

where $\rho \approx 0.2$ for an ideal filter, the amplitude and frequency variations of the resultant signal will be slow, and the filter which defines the feedback-loop selectivity will follow the amplitude-limited resultant of the two signals in a quasi-stationary manner. Consequently, the problem can be solved by considering the two applied signals as equivalent to one signal of slowly varying amplitude and frequency. The analysis we have presented for a single sinusoid applies on a quasi-static basis, and the conclusions reached can be quoted with modifications that take into account the amplitude variations of the signal in the present problem. In particular, we note that the condition for a "feedback steady state" becomes

$$K_s \sin \phi_{fb} < 1 - a \tag{20}$$

and the condition for the feedback-loop gain in the steady state to be less than unity at all instants of time during a frequency-difference cycle is

$$2 K_s \sin (\phi_{fb}/2) < 1 - a. \tag{21}$$

The remarks that were made in connection with conditions (17) and (18) in Section V apply here also. From a comparison of conditions (20) and (21) with the corresponding conditions (17) and (18) that apply in the absence of interference, we conclude that the introduction of the interfering signal will narrow down the locking frequency range and raise the locking amplitude threshold. In a poorly designed system these effects can destroy the reception of the desired signal during those parts of its modulation cycle in which its instantaneous frequency places it where the feedback phase shift is too high. These interruptions will rarely occur during intervals of low-modulation level. The time intervals during which

the interruptions occur generally depend upon the variation of the feedback phase shift with frequency and upon the time function that describes the instantaneous frequency difference between the the two signals. Ideally, the feedback phase shift should be zero over the entire instantaneous frequency range of the desired signal. In a practical system this is difficult to achieve, although reasonable approximations to it can be realized in simple circuits. In any case, direct control over the amplitude of the self-oscillation will enable the operator to decrease the locking threshold by a sufficient amount to save the signal under adverse conditions, with a definite improvement in the capture performance of the receiver.

Let us next consider the situation in which the frequency difference r exceeds $(BW)_{if}/2$. Here, after the opening of the switch S_{if}, three sinusoids will be present across the input of the limiter: the source signals having amplitudes E_s and aE_s at p and p+r rad/sec, and the self-oscillation sinusoid having amplitude kG_{fb} and the frequency ω_o at which the phase shift around the loop is exactly zero. The problem could, of course, be handled in terms of all three sinusoids by tracing several rounds of signal flow around the loop, but the analysis and computation are greatly simplified, if, instead of applying the two source signals simultaneously at the start, we switch on one of them first, allow it to pull the oscillation energy to its own frequency, and then switch on the other signal.

Suppose we switch on the signal of amplitude E_s at p rad/sec, first. If ϕ_{fb} is considered negligible, the applied signal will pull the energy in the self-oscillation to its own frequency, and in the steady state the amplitude of the resultant signal at the limiter input will be given by $E_s + kG_{fb} = E_s(1 + K_s)$. After this locking of the oscillation has been achieved, let us switch on the signal at p+r rad/sec whose amplitude is given by aE_s, a being less than one. The initial ratio of weaker-to-stronger signal, as seen by the limiter, is given by

$$\alpha_1 = \frac{a}{1 + K_s}.$$

At the end of the first round around the loop, the amplitudes of the signals as seen by the limiter will therefore become

$$E_s[1 + K_s A_o(\alpha_1)] \quad \text{at p rad/sec}$$

and

$$E_s[a + K_s A_{-1}(\alpha_1)] \quad \text{at p + r rad/sec,}$$

and the amplitude ratio is now given by

$$\alpha_2 = \frac{a + K_s A_{-1}(\alpha_1)}{1 + K_s A_o(\alpha_1)}.$$

The second and third rounds may also be similarly traced, and so on. Evidently, the steady state will be reached when the resultant ratio of weaker-to-stronger-signal amplitude becomes

$$a_\ell = \frac{a + K_s A_{-1}(a_\ell)}{1 + K_s A_o(a_\ell)}. \tag{22}$$

The weaker signal is thus seen to gain in amplitude at the expense of the stronger by taking away some of the oscillation energy originally captured by this stronger signal. If the balance of oscillation energy is now added up, a part of it will appear to be unaccounted for. It has been split up among components whose frequencies were assumed to fall outside the feedback passband.

Let us next start by switching on the weaker signal at p+r, and allow it to build up its amplitude, at the limiter input, to the value $E_s(a + K_s)$ by pulling the self-oscillation energy to the frequency p+r. If the stronger signal at p is next switched on, two initial possibilities exist. The initial amplitude ratio is given by $a + K_s$. This ratio may either be less than 1 or greater than 1. The case in which this ratio is 1 may be treated as a special case of either of the stated possibilities.

Starting with $a + K_s = \alpha_1$ less than unity, the amplitude ratio at the end of the first round will be

$$\alpha_2 = \frac{a + K_s A_{-1}(\alpha_1)}{1 + K_s A_o(\alpha_1)},$$

and it is easily seen that this ratio will settle

down to the value given by Eq. (22), in the steady state. This conclusion is also readily seen to apply for $\alpha_1 = 1$.

If $a + K_s$ is greater than 1, let

$$\alpha_1 = \frac{1}{a + K_s}.$$

At the end of the first round, the ratio of signal amplitude at p to that at p+r is

$$\alpha_2 = \frac{1 + K_s A_{-1}(\alpha_1)}{a + K_s A_o(\alpha_1)}$$

which is obviously larger than α_1. The second round will, therefore, bring back a larger addition to the amplitude at p, and a smaller one to that at p+r. This follows from the fact that $A_{-1}(\alpha)$ increases, while $A_o(\alpha)$ decreases with increasing α. The ratio at the end of the second round will, therefore, be larger than that at the end of the first round. This ratio will, therefore, build up as we trace the signal flow over successive rounds. At the end of the n^{th} round,

$$\alpha_{n+1} = \frac{1 + K_s A_{-1}(\alpha_n)}{a + K_s A_o(\alpha_n)},$$

and α_{n+1} will be larger than α_n. This build-up will continue until the signal at p, as seen by the limiter, becomes the stronger of the two. If α_{n+1} is greater than unity, while α_n is less than or equal to unity, the argument continues in the $(n+1)^{th}$ round with $A_o(1/\alpha_{n+1})$ at p rad/sec, and $A_{-1}(1/\alpha_{n+1})$ at p+r. In the steady state, the ratio of signal amplitude at p+r to that at p will be that given by Eq. (22).

If the feedback phase shift off the center frequency cannot be neglected, and if $r > (BW)_{if}/2$, the condition for a feedback steady state in which the signals are properly locked is

$$K_s A_o(a_\ell) \sin \phi_{fb} \leq 1,$$

where a_ℓ is the interference ratio at the limiter input in the stable steady state. A corresponding condition for the resultant signal amplitude at the limiter input to exceed $k G_{fb}$ at all times can be readily formulated but the resulting expression is quite involved.

Another interesting situation that can be considered is one in which the undesired signal lies at the frequency of zero phase shift, while the desired signal lies near one of the edges of the passband where ϕ_{fb} is high. Consideration of this situation, and of other possibilities, indicates that conditions (20) and (21) are the most stringent conditions on K_s and ϕ_{fb} that will be encountered when cochannel interference is present.

VII. Cascading

Consider the two-stage cascade illustrated in Fig. 7. When the frequency difference between the two signals exceeds $(BW)_{if}/2$, band-limited positive feedback around the first limiter transforms the interference ratio $\underline{a}$ into a smaller ratio $a_{\ell 1}$ at the input of the first limiter. Narrow-band limiter action in the first stage transforms $a_{\ell 1}$ into a still smaller ratio, a_2, at the output of the first-limiter filter. Positive narrow-band feedback around the second stage reduces a_2 to the ratio $a_{\ell 2}$, which, in turn, is reduced by the second narrow-band limiter to the ratio a_3, and so forth. The various ratios, starting with $a_1 = a$, are related by

$$a_{\ell n} = \frac{a_n + K_{sn} A_{-1}(a_{\ell n})}{1 + K_{sn} A_o(a_{\ell n})}, \quad n = 1, 2, 3 \ldots \quad (23)$$

and

$$a_{n+1} = A_{-1}(a_{\ell n}) / A_o(a_{\ell n}). \quad (24)$$

If, in response to a sinusoidal excitation, the amplitude of the voltage that is fed back to the input of the n^{th} stage is $k_n G_n$, then

$$K_{s1} = k_1 G_1 / E_s$$

and

$$K_{sn} = \frac{k_n G_n}{k_{n-1} A_o\left(a_{\ell(n-1)}\right)}, \quad n = 2,3, \ldots . \quad (25)$$

Since $A_o(a)$ varies slowly between $A_o(0) = 1$ and $A_o(1) = 2/\pi = 0.637$ as $\underline{a}$ varies from zero to one, the ratio K_{sn} will vary over a sufficiently small range (with the small variations of $a_{\ell(n-1)}$) to be essentially a constant. This

indicates that the degree of improvement in the capture performance that is achievable by each of the stages that follow the first is quite independent of the signal level at the output of the i-f amplifier. Only the performance of the first stage will depend markedly upon the signal amplitude delivered by the i-f amplifier. This dependence can best be alleviated by driving this first stage through a narrow-band limiter. Note that such a limiter can be designed to drive the first oscillating-limiter stage at a signal level that exceeds the locking threshold over an entire desired locking-frequency range, and for a wide range of i-f signal levels. Uniformity in the degree of capture improvement, and relative independence of the degree of over-all improvement from the signal level delivered by the i-f amplifier, can thus be secured by cushioning off the effect of variations in the signal amplitudes by means of one or two stages of ideal limiting each of which may or may not have its own positive feedback arrangement. The K_s values of the succeeding stages then become design factors controlled chiefly by the designer to meet prescribed capture performance patterns that follow specifically chosen K_s curves for the various stages.

In Fig. 8 we present curves of output interference ratio versus input interference ratio (when $r > (BW)_{if}/2$ and all bandwidths equal $(BW)_{if}$) for the purpose of comparing the weaker-signal suppression performance of a number of limiters in cascade, each of which has its own regenerative feedback arrangement, with that of a chain of the same number of limiters for which the feedback is taken from the output of the last limiter to the input of the first. The solid and the dashed curves in Fig. 8 describe the relation of a_{out} to $\underline{a}$ for two cascaded limiters with separate feedback branches, for the choices of K_{s1} and K_{s2} indicated on the curves. The dotted curves are for two narrow-band limiters in cascade with feedback taken from the output of the second to the input of the first.

VIII. Interstation-Noise Squelch

When no signal is passed by the i-f amplifier, the self-oscillation at the frequency of zero in-phase feedback will attain an amplitude $k G_{fb} > E_{th}$, and thus the limiter is saturated. In the presence of the random noise that will continue to pass through the i-f amplifier, the oscillation will be sustained as long as $k G_{fb}$ exceeds some threshold value that varies directly with the square root of the i-f bandwidth when the noise is white. The noise component whose frequency is the frequency of zero in-phase feedback is the only component that will encounter a zero locking threshold, and it may well have sparked the self-oscillation. Components off the center frequency will encounter locking thresholds that depend upon the value of ϕ_{fb} at the frequency of each, and, on the average, their amplitudes will be far below these thresholds. Consequently, the situation is one in which a strong oscillation is disturbed by weak noise, and it is equivalent to the situation in which a sinusoid is embedded in low-level noise. The degree of quieting that results from the presence of the oscillation is therefore predictable on the basis of the familiar analysis of low-level FM noise. The noise delivered by the i-f amplifier modulates the strong self-oscillation to produce an FM noise with a parabolic power-density spectrum. Thus, after detection and lowpass filtering, an appreciable degree of interstation-noise quieting can be expected.

When the feedback voltage is decreased to the quieting threshold the noise begins to overpower the oscillation and causes impulsive crackles at the output. As the oscillation amplitude is decreased below the threshold, the noise suppresses the oscillation and breaks through to the output with a softer crackle.

IX. Some Experimental Results

At least four experimental receivers that incorporated limiters with regenerative feedback have thus far been investigated.[5, 6, 7, 8]

In the first receiver, A. W. Emmons constructed a typical low-quality FM receiver with two pentode limiters and applied feedback through

an amplifier around the two limiters. Emmons was able to confirm the theoretical predictions regarding positive versus inverse feedback, the effect on interstation noise, and the improvement in limiter operation.[1,2,3] Within the limitations of his circuit, Emmons was able to realize capture-ratio improvements of approximately 3 to 4 db (receiver capture ratio without feedback was approximately 0.3). For signal levels below the quieting threshold, automatic noise squelch of from 20 to 30 db of quieting was attainable upon application of regenerative feedback.

In a subsequent investigation, L. D. Shapiro evaluated the Emmons receiver, and investigated a new circuit in which the feedback was applied around one pentode limiter through an amplifier. The oscillating-limiter unit was driven from one pentode limiter in order to free its performance from dependence upon the input-signal level and to lower the locking threshold, as seen from the receiver input. Shapiro's results, again, confirmed the theoretical predictions, and many oscillographic illustrations of the performance were accumulated. Figure 9, for example, shows the interstation-noise output from the receiver before and after the application of feedback. Direct measurement showed that approximately 24 db of quieting is represented in this figure.

As part of a detailed study of various topics in FM receiver design, R. J. McLaughlin constructed a receiver in which the feedback was applied through two amplifiers in cascade around one 6BN6 limiter. In the design of this arrangement McLaughlin took notice of the hardships encountered by Shapiro as a result of intolerable amounts of phase shift off the center frequency. The purpose of the two amplifiers in the feedback path was primarily to provide increased feedback gain to counterbalance the low signal level at the output of a 6BN6 limiter and the high attenuation that was introduced by a phase-compensating circuit. The over-all feedback filter consisted essentially of a pair of critically coupled tuned circuits. McLaughlin's results are illustrated in the capture characteristics of Fig. 10 and in the oscillograms of Fig. 11. These results represent the performance of the system when the feedback bandwidth (between half-power points) is 2.3 times the peak-to-peak deviations of the applied signal frequencies. At every measured point on the curve, the feedback was adjusted so that K_s attained its locking-threshold value (in the presence of the interference) for a peak deviation of 50 kc from the center frequency. Note that the oscillograms of Fig. 11 illustrate the quality of the output waveforms that were observed with $a = 0.8$ and $a = 1.4$ during the measurements that led to Fig. 10.

A direct attempt was finally made to synthesize a feedback filter by starting from a specification of the desired form for the feedback phase characteristic. The phase shift should ideally be close to zero within the desired passband, and should rise sharply outside the passband. In view of the limitations that are imposed by non-zero feedback phase shift upon the signals that can pass through the oscillating limiter, it is evident that the ideal phase characteristic would enable the oscillating limiter to simulate an ideal dynamic filter for FM signals. Such a filter would discriminate against incoming signals on the basis of the feedback phase shift at their frequencies and on the basis of their amplitudes as compared with the amplitude of the self-oscillation. Excellent discrimination against adjacent-channel signals would then be achievable, even when these signals are several times stronger than the desired in-band signal. Unfortunately, the amplitude characteristic that accompanies the desired phase characteristic has two extremely sharp peaks at the edges of the band, and dips to uncomfortably low levels within the passband. The effect of the "rabbit ears" (name by courtesy of Professor E. A. Guillemin) at the edges of the band is drastically cut down by the attendant high phase shift in those regions, but the low level of the passband transmission sets a limitation on the feedback signal level.

The simplest realization of a filter whose characteristics approach those just described consists of a pair of overcoupled tuned circuits. But

in an interesting application of modern filter theory, K. P. Luke synthesized a filter that required a total of 7 pentodes (no reflection on modern filter theory). The theoretical and the measured characteristics of this filter are shown in Fig. 12. The phase characteristic was measured by using a method that was suggested by this writer. This method is based on the use of condition (17). Briefly, a sinusoidally modulated FM signal is applied at the center of the band and the feedback is set for a specific value of K_s. The maximun frequency deviation of the signal is then increased until the signal begins to run into the limits of its locking range at the peaks of the modulation. If this happens at a deviation of Δf from the center frequency f_o, the phase shift can be said to be given by $\phi = \sin^{-1}(1/K_s)$ at $f_o + \Delta f$.

In Luke's system, the feedback was taken around three very wideband limiters, each of which utilized a pair of back-biased clipping diodes (the limiter-discriminator system was inherited from a previous experimental study of FM transients). Thus, the three limiters were equivalent to one excellent wideband limiter.

Figure 13 presents oscillograms that describe the effect of regenerative feedback around the limiter upon the i-f selectivity (as seen at the input of the oscillating limiter), the combined i-f and oscillating-limiter selectivity, and the feedback-filter amplitude response.

Figure 14 shows the effect of feedback upon the FM disturbance that results from the interaction between two sinusoids of constant frequencies, when the frequency difference assumes the values specified in this figure.

Figures 15 and 16 present the results of measurements of the locking range and locking threshold as functions of $\underline{a}$ and of the other parameters indicated. In order to understand the behavior of these curves, note that analytical expressions can be derived from condition (20) for the locking range and the locking threshold. Thus, at either threshold,

$$K_s \sin \phi_{fb} = 1 - a. \tag{26}$$

The locking-threshold signal amplitude can be written as

$$E_{LT} = k G_{fb} \sin \phi_{fb}/(1-a), \tag{27}$$

which shows that for fixed feedback voltage, $k G_{fb}$, and fixed locking frequency range (which fixes ϕ_{fb}), E_{LT} = (constant)/(1-a). This dependence of E_{LT} upon $\underline{a}$ is clearly brought out by the experimental curves with good agreement within the limits of the poor experimental accuracy that could be secured in the various measurements that were necessary.

As for the locking frequency range, let us first note from Fig. 12 that ϕ_{fb} varies approximately linearly with frequency deviation from the center frequency over most of the desired passband. Thus we can approximate ϕ_{fb} by $\phi_{fb}(\omega) \approx t_d(\omega - \omega_o)$ within almost all of the desired passband. Thus if $\omega = \omega_\ell$ denotes the frequency at one edge of the locking range, and if LR denotes "locking frequency range," we have $\omega_\ell - \omega_o \approx \phi_{fb}(\omega_\ell)/t_d$ and

$$LR \approx \frac{2}{t_d} \sin^{-1}\left(\frac{1-a}{K_s}\right) \tag{28}$$

$$\approx \frac{2}{t_d K_s}(1-a),$$

where the approximation holds for $(1-a)/K_s \leq 0.775$ and within the limitations of the assumption of linear phase dependence upon $\omega-\omega_o$. Again, the agreement between the experimental curves and these expressions is good, as a few numerical checks will convince the reader.

Finally, the oscillograms of Figs. 17 and 18 show an extremely interesting result. Under the conditions of operation specified in these figures, the application of regenerative feedback is seen, from Fig. 18, to pull an FM signal from below the level of readability that is encountered in a conventional FM receiver. In order to understand the reason for this, the oscillograms of Fig. 17, which correspond in their arrangement to those in Fig. 18, are presented. In the study of these oscillograms, it is helpful to remember that the feedback phase shift (see Fig. 12a) is negligible over the frequency range covered by the

instantaneous frequency of the signal. Consequently, the locking threshold at the frequencies of peak deviation is set by $K_s \approx 15$.

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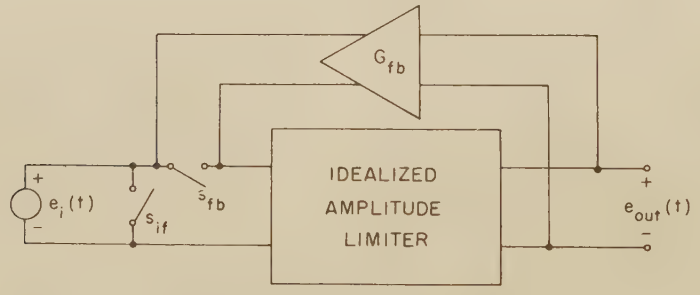


Fig. 1. Limiter with feedback at the intermediate frequency.

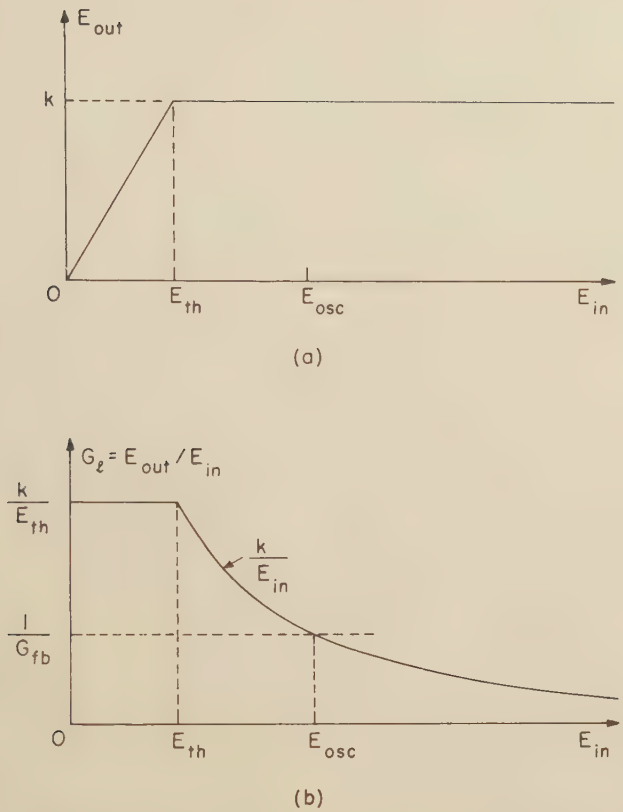


Fig. 2. Characteristics of the limiter: (a) idealized output-vs.-input sinusoidal amplitude characteristic and (b) idealized sinusoidal-amplitude amplification characteristic.

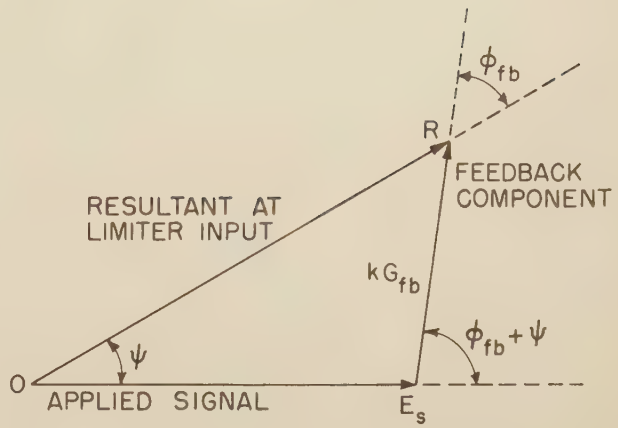


Fig. 4. Signal-phasor relationships at the input of the limiter after the energy in the self-oscillation has been transferred to the frequency of the applied signal.

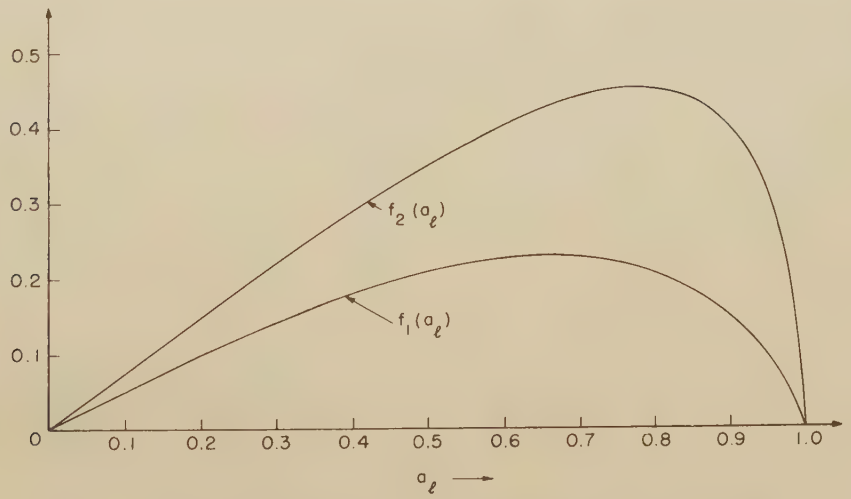


Fig. 3. Normalized reduction in the interference ratio at the input of the first limiter, caused by regenerative feedback under the condition of most severe feedback band limitation. The curve marked $f_1(a_\ell)$ is for feedback around one limiter; $f_2(a_\ell)$ is for feedback around two cascaded limiters. [See Eq. (7).]

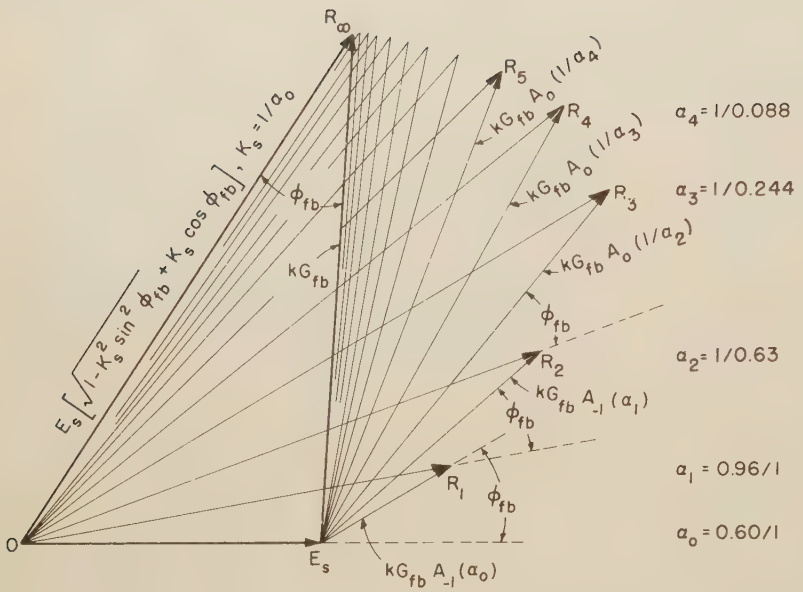


Fig. 5. Pattern of evolution at the limiter input of the resultant component at the frequency of the applied signal. The spurious component at p + 2r rad/sec was ignored in the computations that were made for the construction of this pattern.

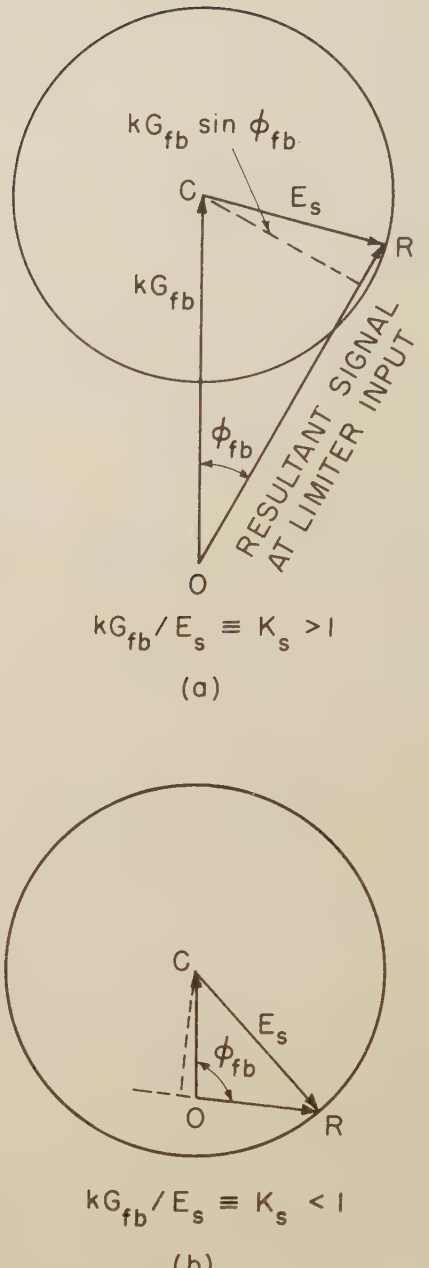


Fig. 6. Phasor constructions that illustrate the significance of condition (18) as compared with condition (17). In these constructions $\overline{OR} = \overline{OC}$, and therefore $\overline{CR} = E_s = 2k\, G_{fb} \sin(\phi_{fb}/2)$. The perpendicular from C to OR represents $k\, G_{fb} \sin \phi_{fb}$.

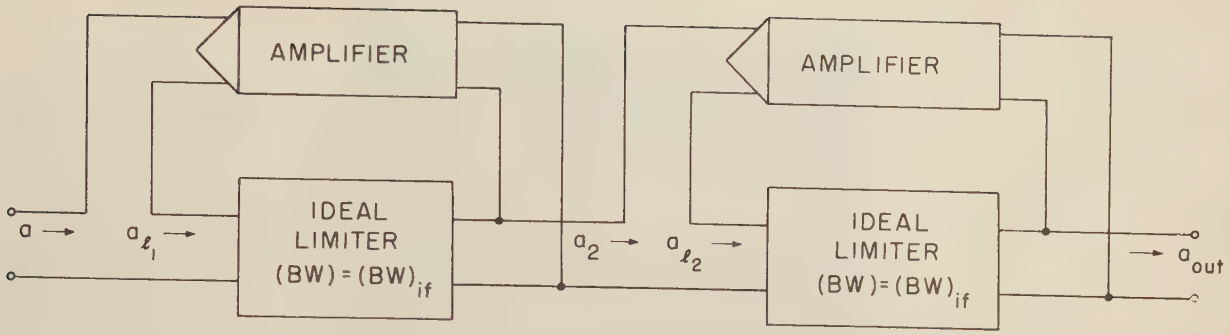


Fig. 7. Two-stage cascade of limiters with independent regenerative feedback branches.

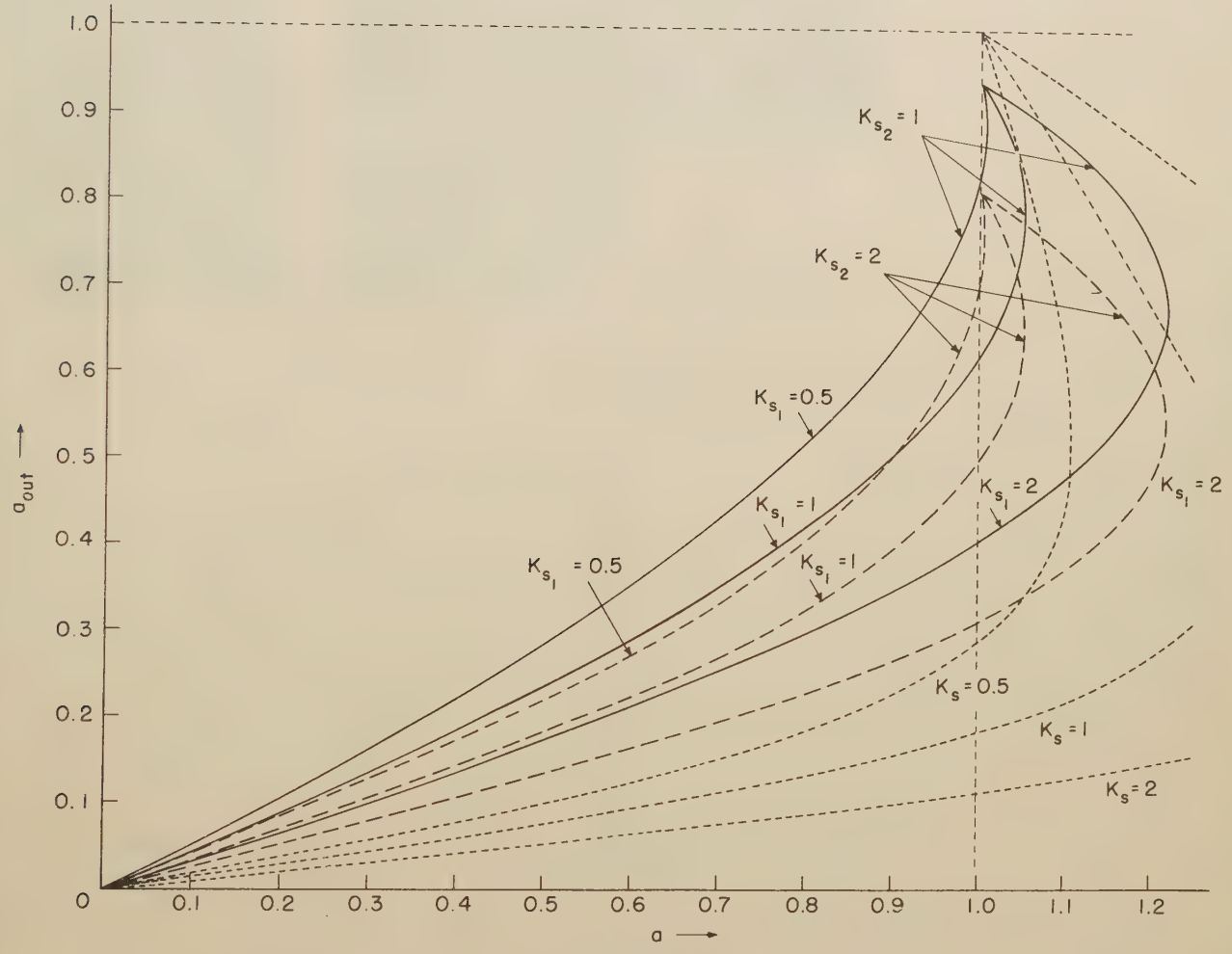


Fig. 8. Output interference ratio vs. input interference ratio (when $r > (BW)_{if}/2$) for two cascaded limiters with independent feedback branches (solid and dashed curves) and for regenerative feedback around two cascaded limiters (dotted curves). All limiter and feedback bandwidths equal $(BW)_{if}$.

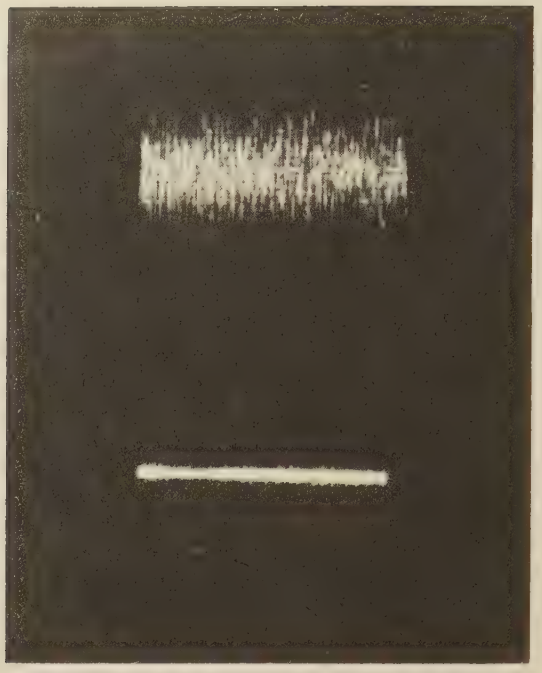


Fig. 9. Automatic interstation-noise squelch. The upper oscillogram records output noise before application of feedback; the lower oscillogram records output noise after application of regenerative feedback. Noise is suppressed approximately 24 db.

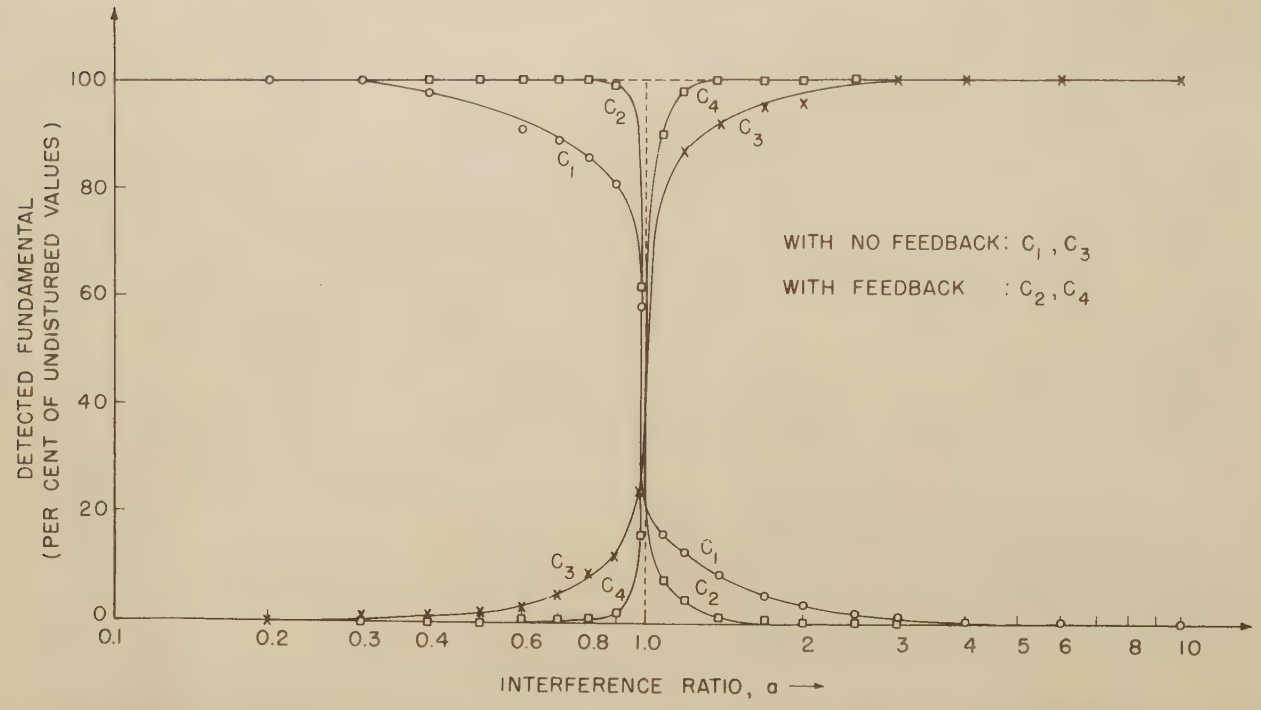


Fig. 10. Effect of regenerative feedback upon capture characteristics of fm receiver. Both signals are centered in the pass band. Each is deviated ± 50 kc. One signal is modulated by 400 cycle tone; the other signal by 1,000 cycle tone. Feedback bandwidth between half-power points is 230 kc. Curves C_1 and C_3 give performance before application of feedback; curves C_2 and C_4 give performance after application of regenerative feedback. At every measured point, K_s is adjusted for its locking-threshold value.

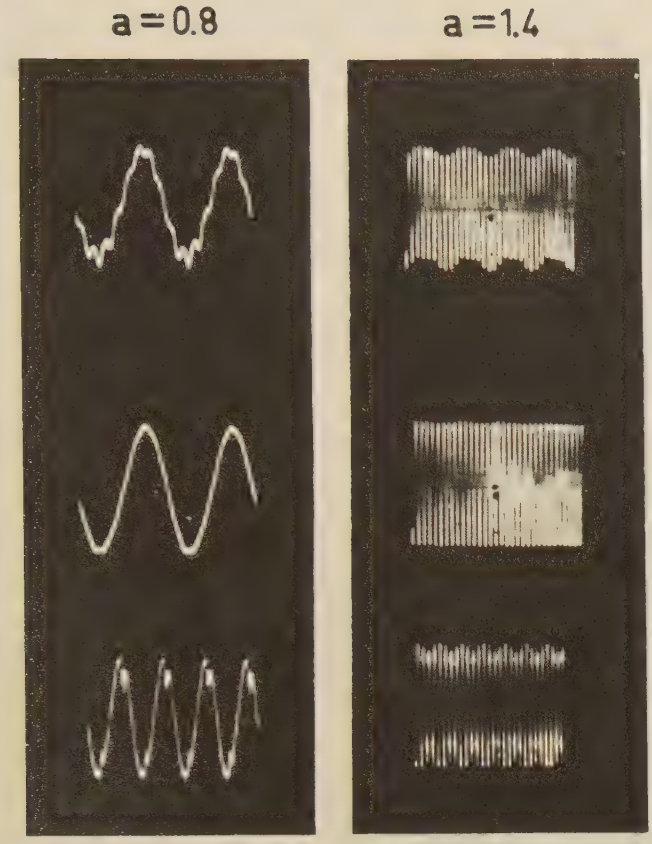


Fig. 11. Output waveforms observed during measurements made for Fig. 10. The top oscillograms record output waveforms before application of feedback; the middle oscillograms record output waveforms after application of regenerative feedback, with K_s set to its locking-threshold value; and the bottom oscillograms record output waveforms after application of regenerative feedback but with K_s slightly exceeding its locking-threshold value.

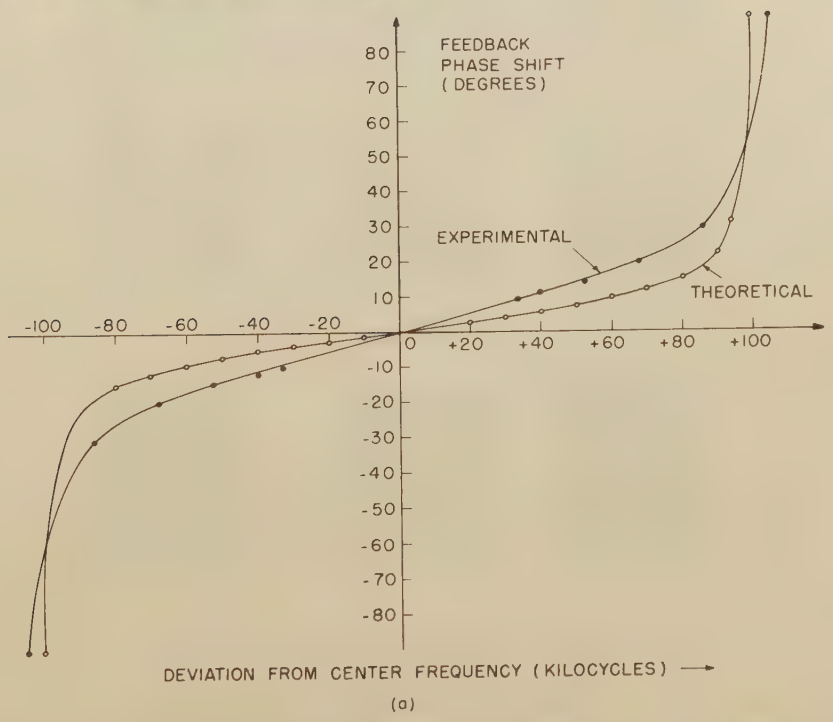
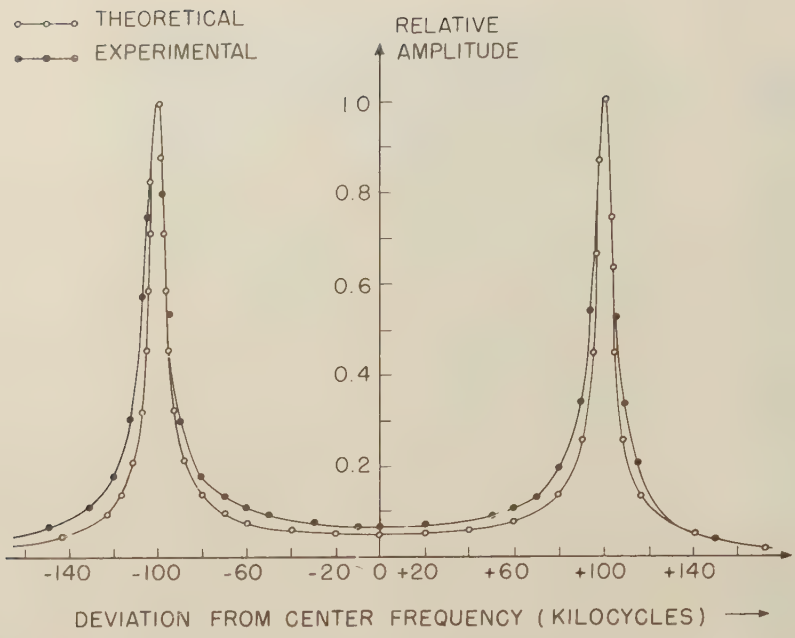


Fig. 12. Theoretical and measured characteristics of a feedback filter that was designed to realize approximately zero phase shift within the desired pass band and sharp rise in phase shift beyond the edges of the pass band.



(b)

Fig. 12 (Continued)

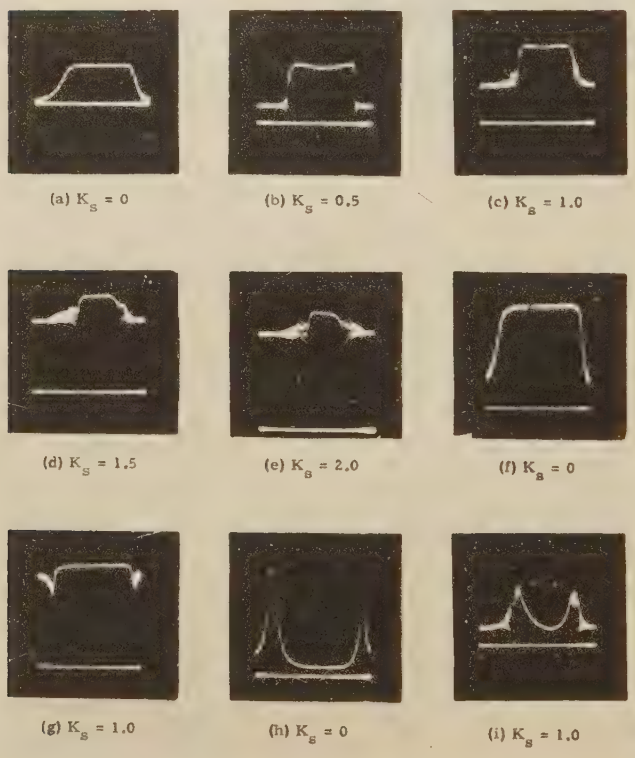


Fig. 13. Effect of regenerative feedback around the limiter upon the quasi-stationary response-amplitude characteristic of (i) IF amplifier as observed at oscillating-limiter input, a, b, c, d, and e; (ii) combination of IF amplifier and limiter (with and without feedback), f and g; and (iii) feedback filter, h and i.

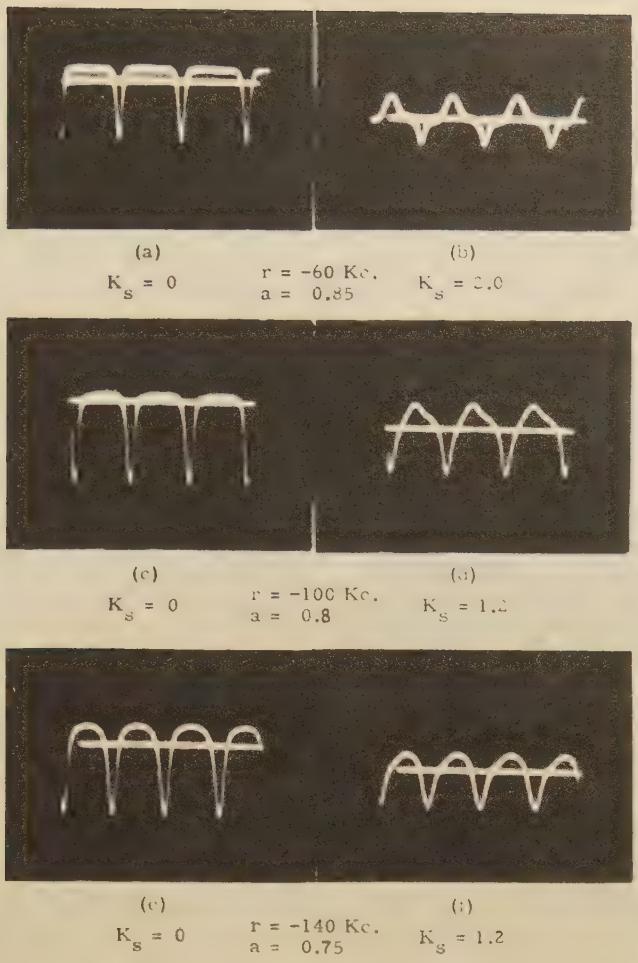


Fig. 14. Effect of regenerative feedback around the limiter upon the fm disturbance (instantaneous-frequency spike pattern) that results from interaction between two sinusoids of fixed frequencies: $E_s \cos pt$ and $aE_s \cos(p + r)t$.

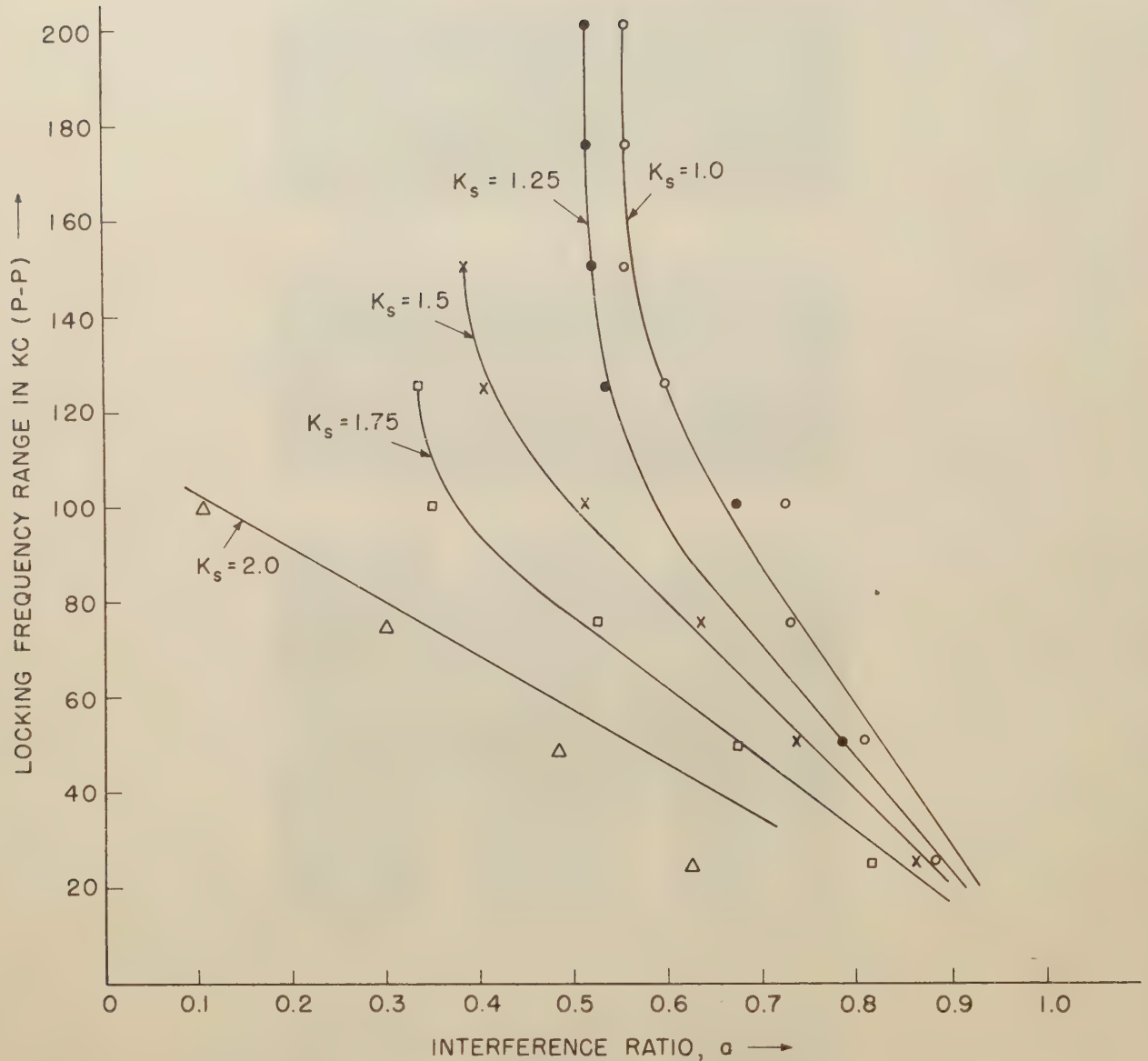


Fig. 15. Variation of measured locking frequency range with the interference ratio, a, for different adjustments of the regenerative feedback around the limiter.

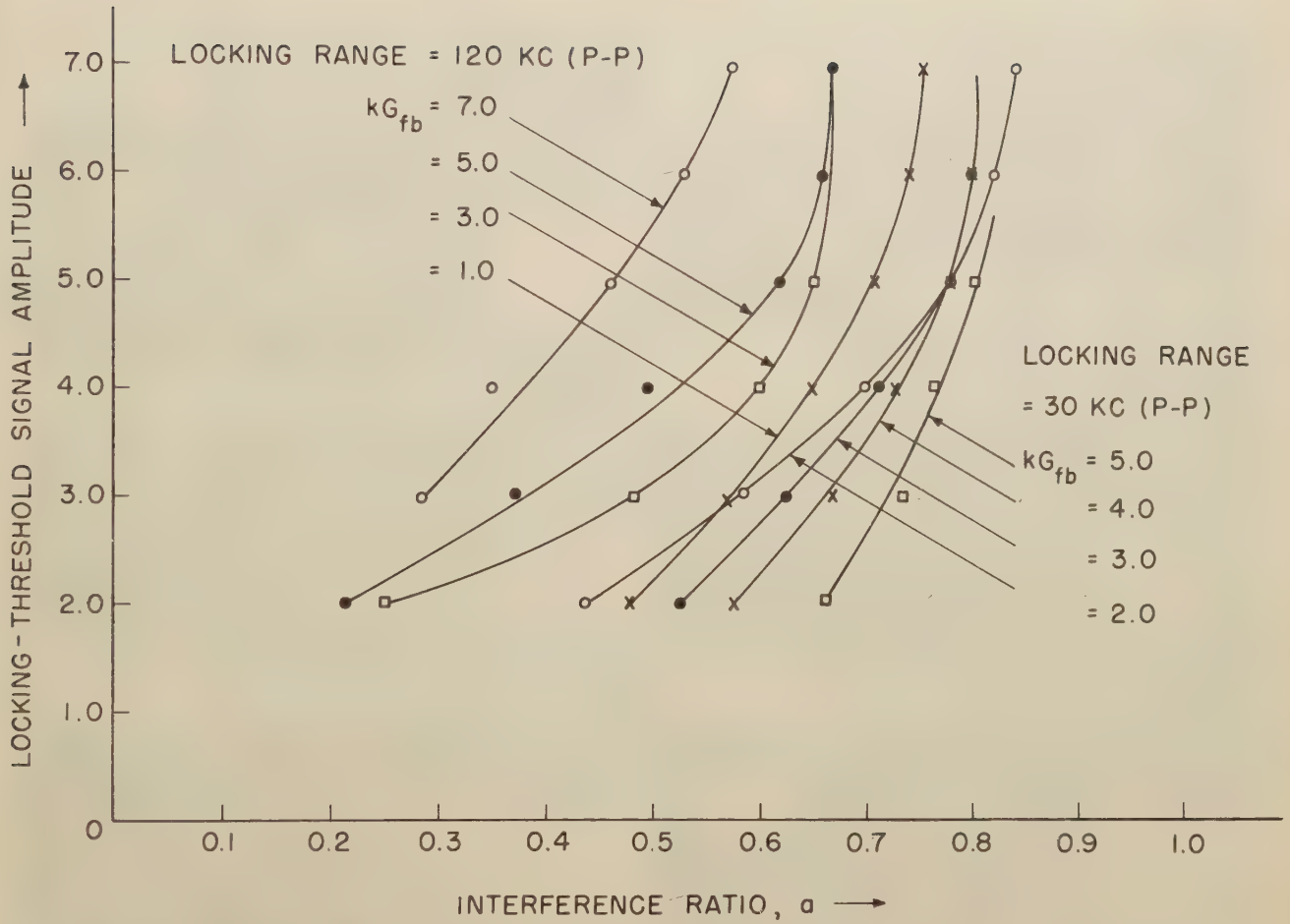


Fig. 16. Variation of measured value of locking threshold with the interference ratio, a, for different adjustments of the regenerative feedback around the limiter.

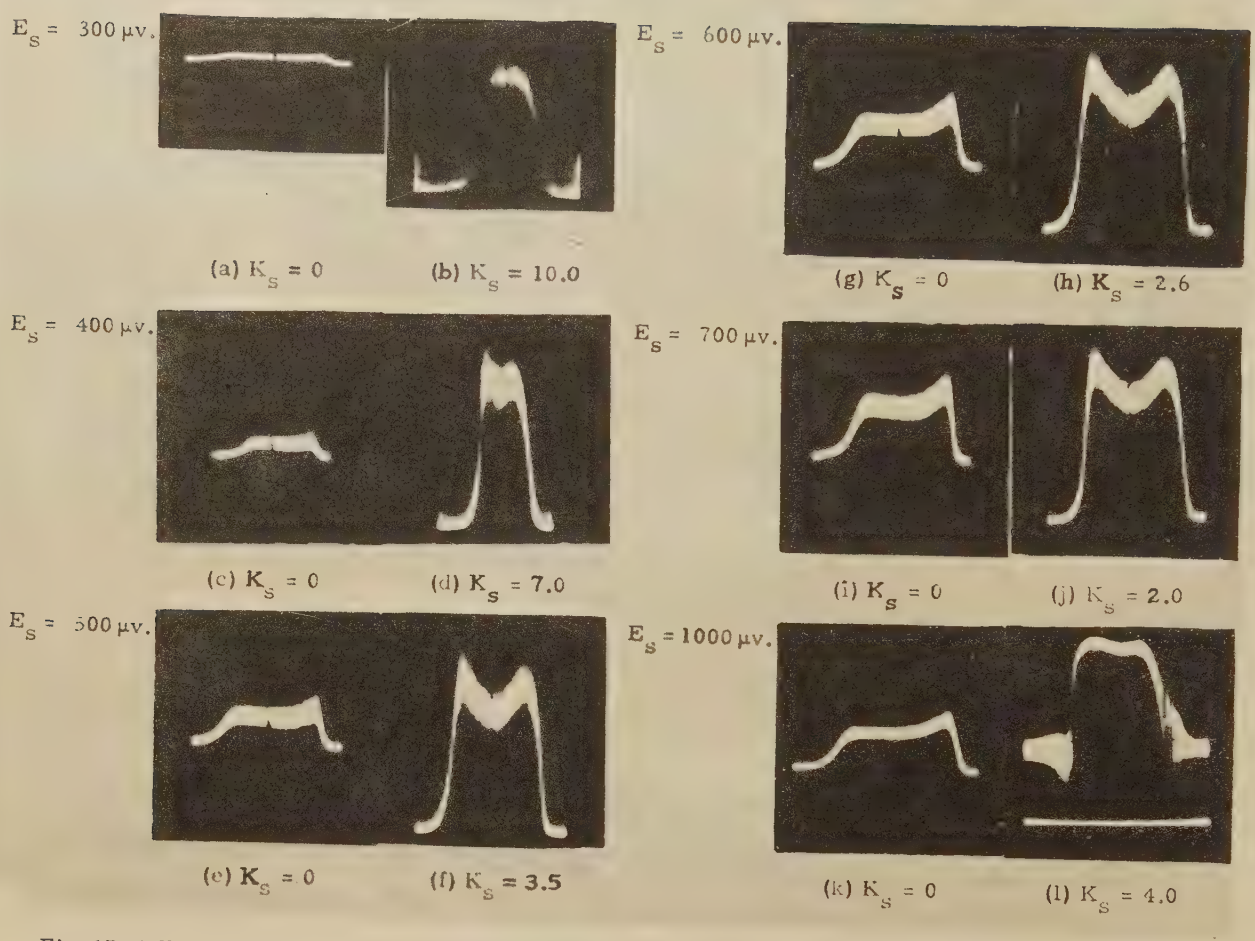


Fig. 17. Effect of regenerative feedback around the limiter on the signal level at the input of the limiter. Here, E_S represents the level of the signal that is applied at the IF amplifier input.

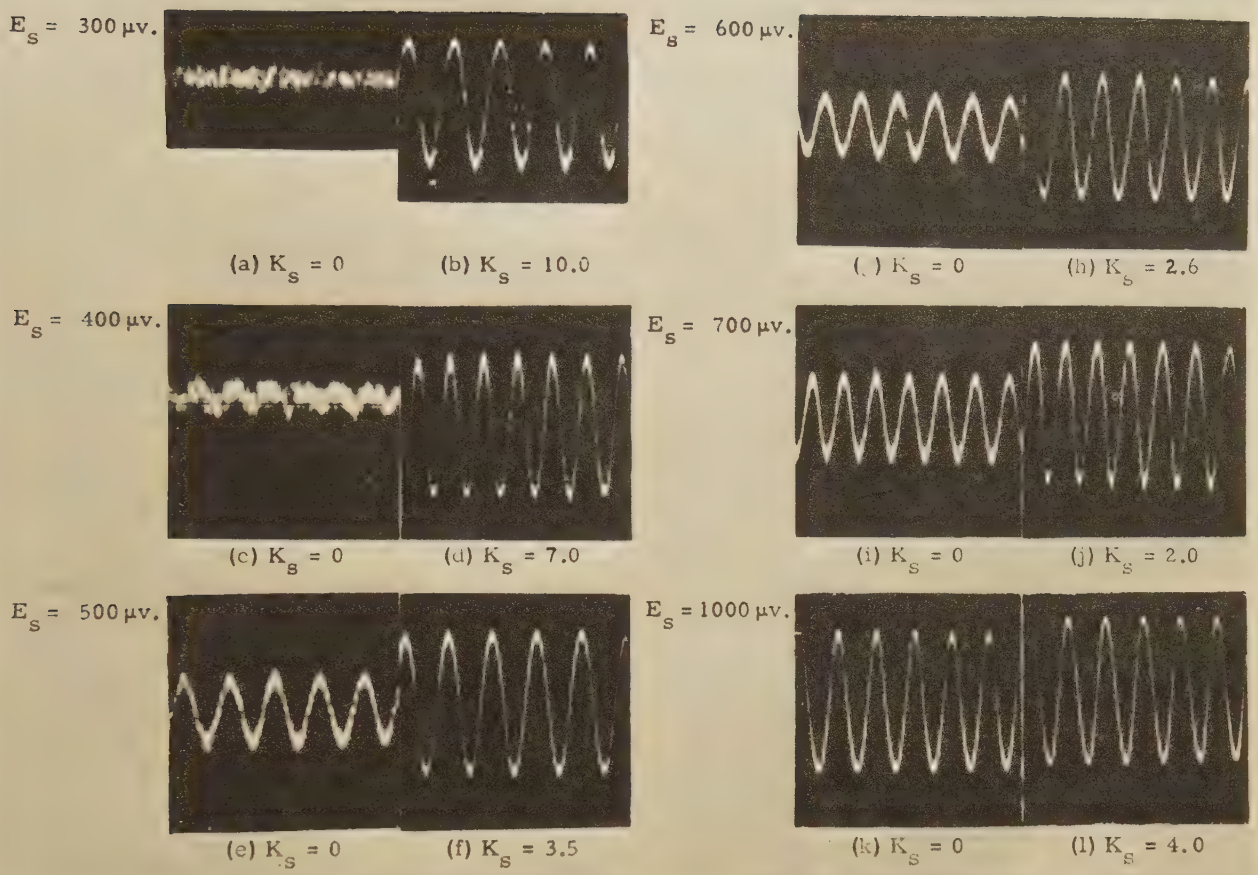


Fig. 18. Effect of regenerative feedback around the limiter upon the detected waveform at the output of the fm receiver under the same conditions that are illustrated in Fig. 17. The fm signal is deviated ±20 kc by a single tone. Notice the marked improvements in "readability" that are revealed, particularly in (a), (b), (c), and (d). In (a) and (c) the signal is almost completely overwhelmed by the ambient noise before the application of feedback; in (b) and (d) the application of regenerative feedback enables the signal to predominate over the noise. The locking-threshold value of K_s for the applied signal is approximately 15.

A NEW APPROACH TO COMPACTNESS IN MOBILE RADIO-TELEPHONE DESIGN

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Summary

A mobile radiotelephone having a power rating of 30 watts is described. The unit incorporates transmitter, receiver, operating controls and loudspeaker in a single compact aluminum alloy case. Total weight is less than twelve pounds. The unit is small enough for underdash mounting in a modern passenger car. The transistor power supply is in a separate case designed to be mounted on the vehicle firewall. Heater voltage regulation is employed to maximize vacuum tube reliability. Careful attention to thermal design has minimized the development of hot spots in the interior of the equipment.

Introduction

There is a continuing trend in the design of civilian Vehicular Communications equipment towards size reduction. The trend is not as striking as in military equipment since the civilian equipment designer must, in general, meet the requirement of reduction in size without increase in cost. Of course, another essential requirment is reduction in size without any loss of equipment reliability.

A typical 150 Mc/s equipment of 1946 vintage had a receiver and transmitter in separate cases, a third case for a dynamotor and junction box and a remote control head. Total volume was in excess of 3800 cubic inches and total weight was 85 lbs. Battery drain on transmit for 20 watts output was 40 amperes at 6 volts or 240 watts. A 150 Mc/s unit of the same manufacture of current production has a volume of 900 cubic inches and a weight of 41 lbs. Battery drain on transmit for 20 to 25 watts output is 33 amperes at 6 volts or 200 watts. This is fairly typical of equipment now going into service.

Vibrator supplies are still common in current production units although transistorized power supplies are coming into ever-increasing usage. Except for power supplies and microphone pre-amplifiers transistors have not yet supplanted the vacuum tube in vehicular units except in a few advanced designs whose cost is considerably higher than conventional units.

The older units were, of course, of necessity, mounted in the trunk in passenger car usage. Today's units normally have an option of trunk or underdash mounting. The latter facility isusually achieved by attaching a control head to the end cover of the mobile case thus achieving single unit construction. However, year by year, unfortunately, the distance between the dashboard and the road surface seems to be decreasing and a case size of the order mentioned above presents very real difficulties in mounting the unit underdash in modern cars.

The objectives of the radiotelephone design descibed in this paper were to produce a unit having a power output of 30 watts at 174 Mc/s which could fulfill the following requirements:

(1) Practical underdash mounting in modern cars.

(2) Cost not to exceed standard designs in present use.

(3) Performance parameters and reliability at least equal to standard designs in present use.

(4) Availability in wide-band or split channel and in single, dual and triple channel models.

(5) Power supply to be transistorized with a transistorized voltage regulator included for stabilizing vacuum tube heaters.

The design as it is now being produced has the main transmitter, receiver and control circuits in a single underdash mounted case. The transistor power supply and the voltage regulator are in a second small box normally mounted on a vehicle firewall. Thus the complete installation is in the vehicle area adjacent to the driver. The separation of the transistor circuits is a precaution to prevent overheating of the transistors by proximity to the higher temperature vacuum tube circuits.

Fig. 1 is a photograph of the SY30 unit illustrating the styling concept employed. Although the photgraph illustrates a non-vented case, the case design finally adopted is vented to meet thermal design requirements. The dimensions of the main unit are 4 x 11 x 12 with a weight of 12 lbs. The power supply dimensions are 7 x 5-1 2 x 2-1 2 with a weight of 4 lbs. Thus the total volume of the two units is only 600 cubic inches and the total weight 16 lbs. The transmit drain is 13 amperes at 13.6 volts or 177 watts input for 30 watts output.

While the equipment has been designed to operate in an ambient of 60°C, in practice the operating conditions with respect to maximum ambient temperature encountered are much more favourable in underdash mounting than in trunk mounting. This applies both to the temperature of the surrounding air and to the possibilities for convection and radiation cooling.

Many current designs employ much larger cases of a dust tight non-vented construction. Venting of the case makes possible a more compact design without increasing internal temperatures. In many years of experience with both vented and non-vented construction it has been noted that the only negative feature of importance in the vented construction has been increased trouble with relay contacts. The SY30 design employs a single relay for both power and antenna switching and houses this relay in a dust proof metal enclosure.

Time limitations prevent a complete discussion of the design and therefore only a number of selected topics which represent areas in which the design has departed from conventional practice will be presented. These topics are multi-channel operation, transmitter power amplifier, receiver I.F. filter and heater voltage regulation.

Due to the fact that the equipment is not required to be remotely operated by a control head, the problem of multi channel operation is considerably simplified. Up to three channel operation is available with crystal switching being done by a wafer switch operated from the front panel. The crystal oven employed for three channel units has been especially designed for this unit. The oven, which mounts three HC-6U crystals, plugs into an octal socket and employs a snap action thermostat for long life and high reliability.

The most interesting aspect of the transmitter design is the power amplifier. Engineering design is always, of course, a question of choosing from amongst many possible alternatives to achieve the best balance between economic considerations and equipment performance. In choosing a tube for the final amplifier of the transmitter it seemed that only two basic types could be considered - a single tetrode in a soft glass envelope and a dual tetrode in a hard glass envelope. A representative example of the first type of tube is the 6146 and of the second type the 6252. Since the dual tetrode has three times the cost of the single tetrode type, the final decision to employ the latter type was reached only after careful consideration of relative features. Incidentally in comparing the two tube types there is no intention to cast slurs on the design of the lower cost tube since it certainly delivers the performance claimed for it by the manufacturer.

The basis of the comparison is the requirement of a power amplifier to deliver 30 watts into a load at 175 Mc/s at a satisfactory level of reliability.

The highest frequency for maximum ratings is 200 Mc/s for the 6252 and 60 Mc/s for the 6146.

At 175 Mc/s the 6252 will deliver 39 watts into a load within CCS ratings and with a plate efficiency of 80%. At the same frequency the maximum CCS rated output of the 6146 is 25 watts with a plate efficiency of 69%. On an ICAS basis a power output of 35 watts is obtainable from a 6146. However, it should be noted that the figures of power output for the 6146 are listed by the manufacturer as "approximate". In practice it is found that at 175 Mc/s the tube to tube variations of power output are considerably greater with the 6146 than with the 6252 due to the fact that the former tube is being operated so far beyond its upper frequency limit for maximum ratings.

The driving power required by the 6252 for 39 watts of output is 1.5 watts whereas the driving power required by the 6146 for 25 watts of output is 3 watts. Again this latter figure is listed as "approximate" and it has been found in practice that the previous comments relative to tube to tube variation of power output also apply to driving power. Thus while it is common practice to drive the 6146 with another similar transmitting type tube, such as a 2E26, in the SY30 design it has been found that a much smaller receiving type tube, the 12BY7, provides adequate drive for 30 watts of RF output.

The 6252 is rated at a maximum bulb temperature of 250°C and the 6146 for a maximum bulb temperature of 220°C. Since the 6252 operates at greater efficiency and has a larger heat dissipating area of glass envelope these various factors combine to give the dual tetrode a definite advantage in a more compact design where problems of heat dissipation must be given more serious consideration than in

a larger unit.

Having decided on the tube types to be employed for the power amplifier and driver stages, a further precaution was taken in the method of setting the operating points for Class C operation. In many Class C Land Mobile transmitter power amplifiers screen grid and control grid operating voltages are determined largely by the current drawn by these elements. When grid leak bias is used control grid voltage is determined by control grid current and when large values of screen dropping resistance is used screen grid voltage is mainly determined by screen grid current. Unfortunately for a given set of electrode voltages the tube to tube spread of control grid and screen grid currents is of the order of three to one. As a result the use of grid leak bias and large values of screen resistance results in pronounced changes of power amplifier output and efficiency when tubes are changed. The SY30 design employs 80% fixed bias and a screen voltage supply of good regulation. As a result a very consistent and reproducible performance has been obtained between transmitters and with different power amplifier tubes in a given transmitter. Of course, an additional advantage of fixed bias is protection of the power amplifier tube in the event of drive failure.

In order to produce a compact receiver I.F. filter design ferrite cup cores are employed with windings of 20/44 Litz. Whereas with powdered iron core materials the maximum attainable Q in the available space is about 120, the use of ferrite cores permits Q's at 455 Kc/s of 250. The coils are permeability tuned with a ferrite slug. Fig. 2 illustrates the dimensions of a filter which is capable of housing up to 10 resonators. A patented method of temperature compensation results in negligible shift of centre frequency so that temperature change merely widens or narrows the filter passband as the Q changes with change in resistance of the winding. Fig. 3 illustrates this point for a particular design of split channel filter.

Fig. 3 illustrates the characteristics of a 10 resonator narrow band filter at temperatures of 25°C, -30° and 85°C. Centre frequency drift relative to room temperature is 450 cycles at 85°C and -500 cycles at -30°C. The 6 db bandwidth is 13.1 Kc/s at 25°C and decreases to 10.2 Kc/s at 85°C and increases to 16.8 Kc/s at -30°C. The 80 db bandwidth is 27.5 Kc/s at 25°C and decreases to 24.0 Kc/s at 85°C and increases to 29.5 Kc/s at -30°C.

Tests were conducted to determine the variation in SINAD sensitivity and selectivity as measured by the two generator test with various filters.

All the filters are built into identical housings as shown in Fig. 2.

In Figs. 4, 5, 6, note the flatness of the pass band and the rounding off of the nose at the cut off frequencies rather than any abrupt discontinuity of the insertion loss characteristic. These are desirable features from the viewpoint of minimizing transient responses resulting from impulse interference.

Fig. 4 shows the characteristic of an 8 resonator wide band filter. The 6 db bandwidth is 38 Kc/s and the 70 db bandwidth is 81 Kc/s. The insertion loss at centre frequency is 4 db.

Fig. 5 shows the characteristic of an 8 resonator split channel filter. The 6 db bandwidth is 19 Kc/s and the 70 db bandwidth is 44 Kc/s. The insertion loss at centre frequency is 6.5 db.

Fig. 6 shows the characteristic of a 10 resonator split channel filter. The 6 db bandwidth is 13.1 Kc/s and the 80 db bandwidth is 27.5 Kc/s. The insertion loss at centre frequency is 10 db.

Sensitivity of the SY30 receiver was 0.34 microvolts for the wideband filter and 0.32 microvolts for both narrowband filters. Deviation was of course adjusted to the appropriate value for the wide and narrow band tests, respectively.

The two signal generator selectivity standards set by EIA are 70 db for both wide and narrow band receivers. The measured results were as follows:

8 Coil wideband filter - 85 db for interfering signal at plus or minus 60 Kc/s.

8 Coil narrowband filter - 78 db for interfering signal at plus or minus 30 Kc/s.

10 Coil narrowband filter - 80 db for interfering signal at plus or minus 30 Kc/s.

Present production of SY30 employs the

appropriate 8 coil filters in both the wideband and split channel models.

It was considered very important from a reliability point of view to allow for heater voltage regulation in the SY30 design. After considerable study of many possible regulator schemes it appeared that all feasible regulators were in one of three categories:

(1) Non linear resistances
(2) Electromechanical devices
(3) Transistor and zener diode circuits.

Non linear resistances were ruled out because of their low efficiency and relatively poor regulating characteristic. Electromechanical devices were ruled out because of cost and problems of restricted mounting position and susceptibility to shock and vibration in other cases. There are indications however that satisfactory electromechanical regulators may be available.

Use of a transistorized voltage regulator was complicated by the fact that under some conditions appreciable power of the order of 20 watts must be dissipated by the transistor under conditions of high ambient temperature. From a cost point of view the only practical transistors available were made in the PNP configuration only. It was therefore necessary to regulate the negative leg of the heater supply as will be apparent from the circuit used. Since most modern cars have the negative side of the battery grounded it was necessary in the SY30 design to remove all heaters from the chassis ground. Moreover, to ensure a conservative design the heater load was split into two sections each taking approximately 2.5 amperes at 12 volts and each regulated by a separate transistor. Fig. 7 illustrates the simplicity of the regulator circuit. The voltage reference is a 10 watt Zener diode selected to 5% accuracy. Initially low gain transistors, 2N442, were employed giving the regulation characteristic shown in Fig. 8. However, by the use of the higher gain 2N278 type, the improved regulation curve shown in Fig. 9 was obtained.

Although the initial summary of this paper referred to the 60 watt version of this unit no description of the 60 watt unit is included in the above paper since development work on this model is still in process.

Acknowledgement is made of the efforts of Messrs. Teosi, Ward and Willmott of the Systcoms Engineering Department who were responsible for the development of the SY30. Acknowledgement is also made of the assistance of Mr. Bernard Tennant who developed the I.F. filter and of Mr. D. S. Simkins who provided valuable application engineering assistance in the use of the 6252 power amplifier tube. Messrs. Tennant and Simkins are associated with the Rogers Electronic Tubes and Components Division of Philips Electronics Industries of Canada.

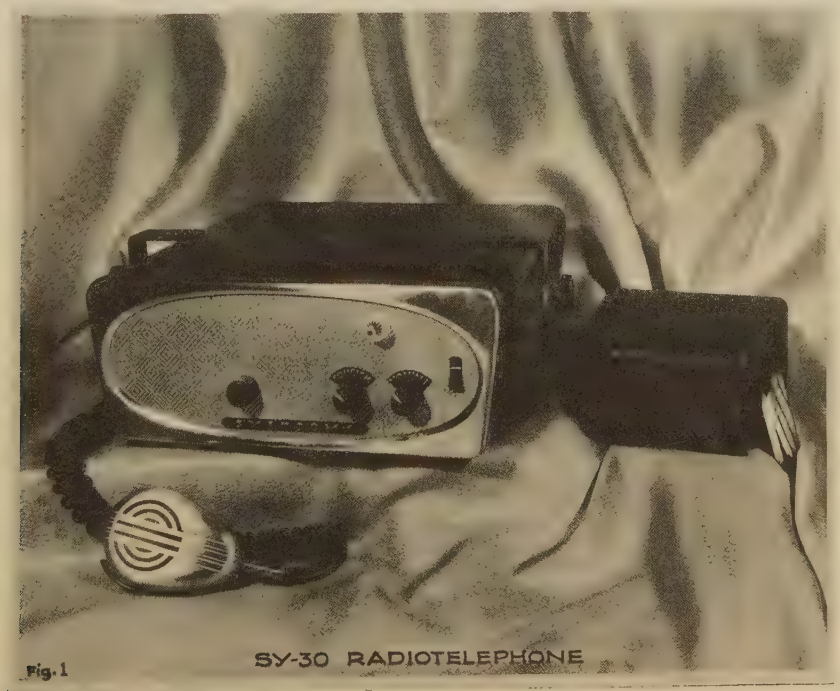


Fig. 1 SY-30 RADIOTELEPHONE

Fig. 2 RECEIVER I.F. FILTER

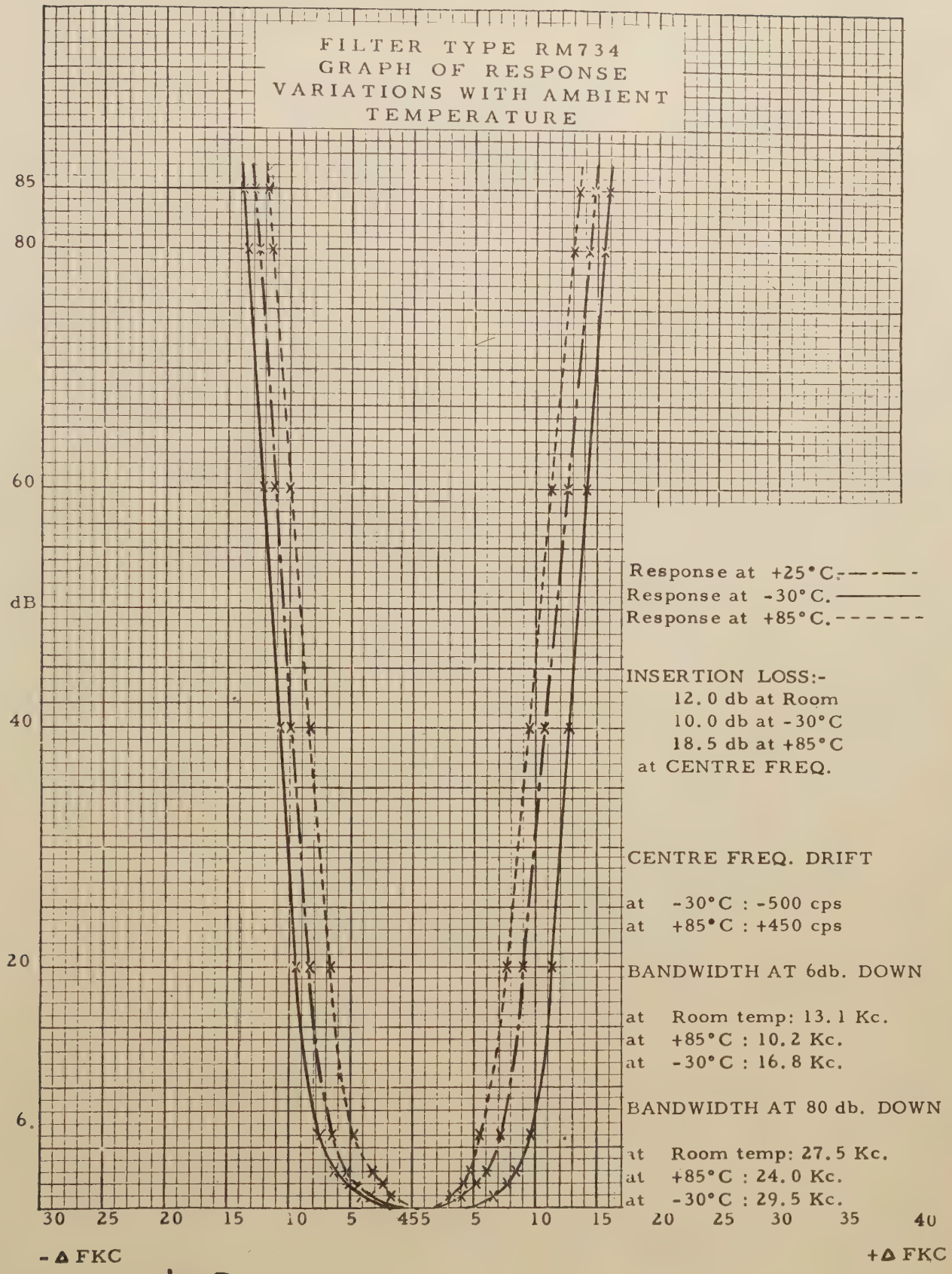

FILTER TYPE RM734
GRAPH OF RESPONSE
VARIATIONS WITH AMBIENT
TEMPERATURE
85
80
60
dB
40
20
6.
Response at +25°C.
Response at -30°C.
Response at +85°C.
INSERTION LOSS:-
12.0 db at Room
10.0 db at -30°C
18.5 db at +85°C
at CENTRE FREQ.
CENTRE FREQ. DRIFT
at -30°C : -500 cps
at +85°C : +450 cps
BANDWIDTH AT 6db. DOWN
at Room temp: 13.1 Kc.
at +85°C : 10.2 Kc.
at -30°C : 16.8 Kc.
BANDWIDTH AT 80 db. DOWN
at Room temp: 27.5 Kc.
at +85°C : 24.0 Kc.
at -30°C : 29.5 Kc.
30 25 20 15 10 5 455 5 10 15 20 25 30 35 40
-Δ FKC
+Δ FKC

Fig. 3

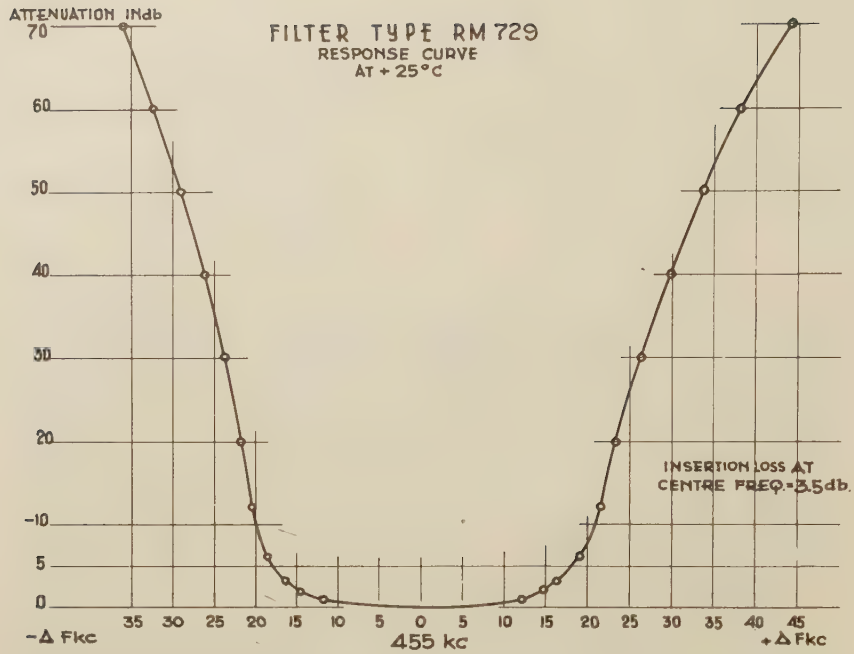

ATTENUATION INdb
FILTER TYPE RM 729
RESPONSE CURVE
AT + 25°C
70
60
50
40
30
20
-10
5
0
35 30 25 20 15 10 5 0 5 10 15 20 25 30 35 40 45
INSERTION LOSS AT
CENTRE FREQ. = 3.5db.
-Δ Fkc
455 kc
+Δ Fkc

Fig. 4

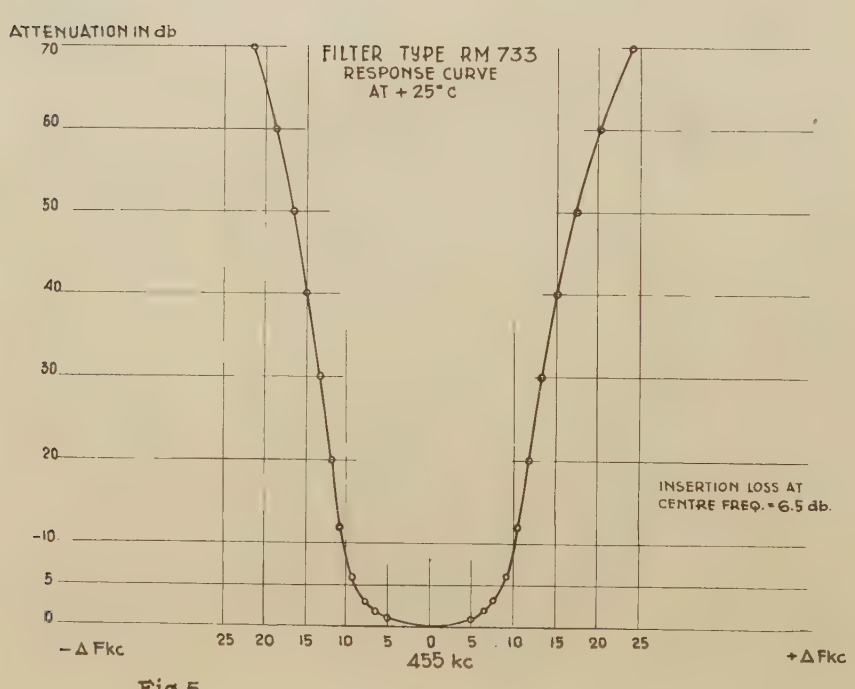

ATTENUATION IN db
FILTER TYPE RM 733
RESPONSE CURVE
AT + 25°C
70
60
50
40
30
20
-10
5
0
25 20 15 10 5 0 5 10 15 20 25
INSERTION LOSS AT
CENTRE FREQ. = 6.5 db.
-Δ Fkc
455 kc
+Δ Fkc

Fig. 5

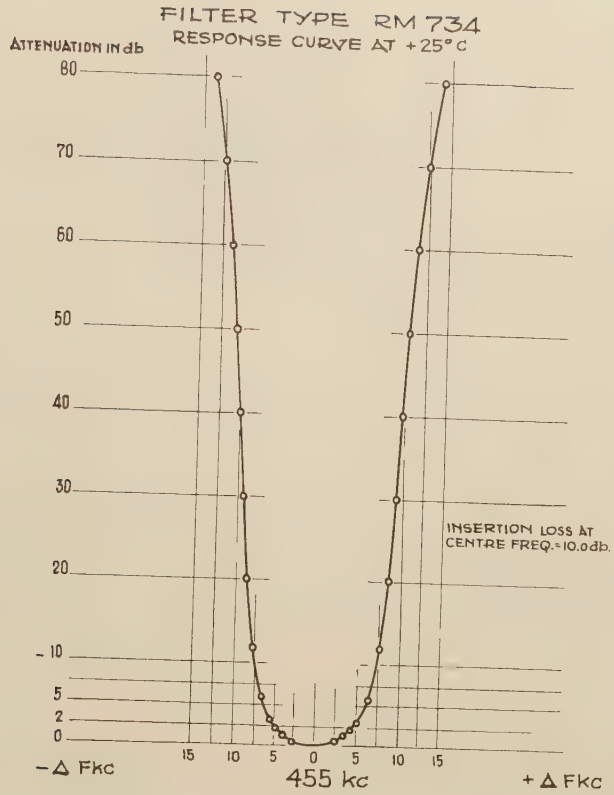
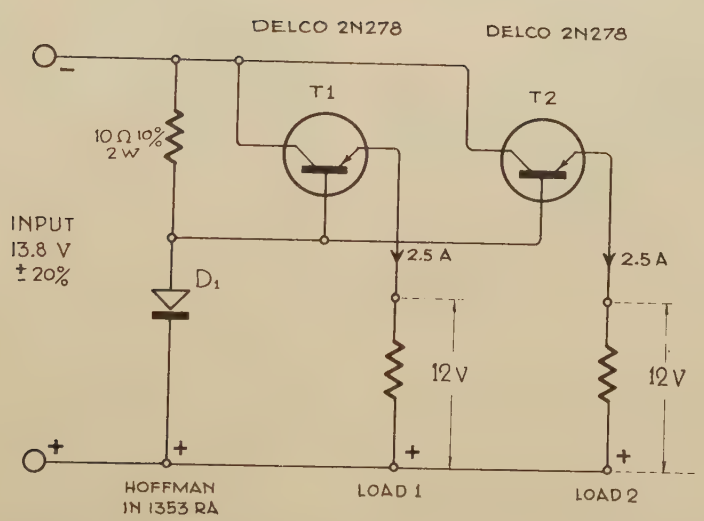
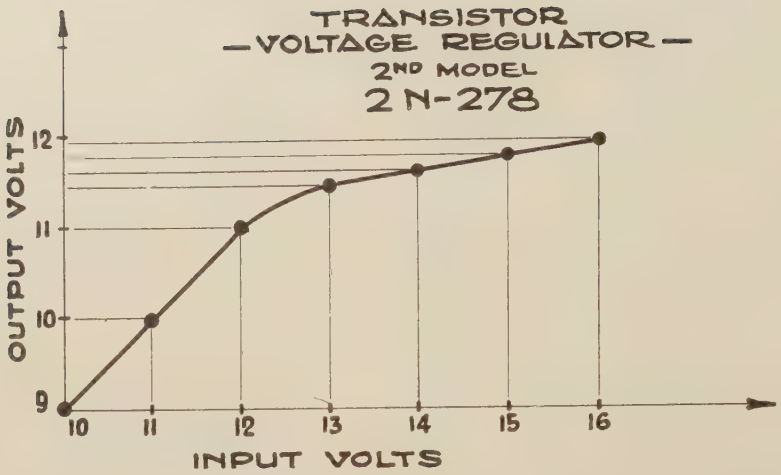


Fig. 6.



TRANSISTORIZED HEATER VOLTAGE REGULATOR

Fig. 7.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—VOLTAGE REGULATOR—
2ND MODEL
2 N-278
OUTPUT VOLTS
12
11
10
9
10
11
12
13
14
15
16
INPUT VOLTS

Fig. 8.

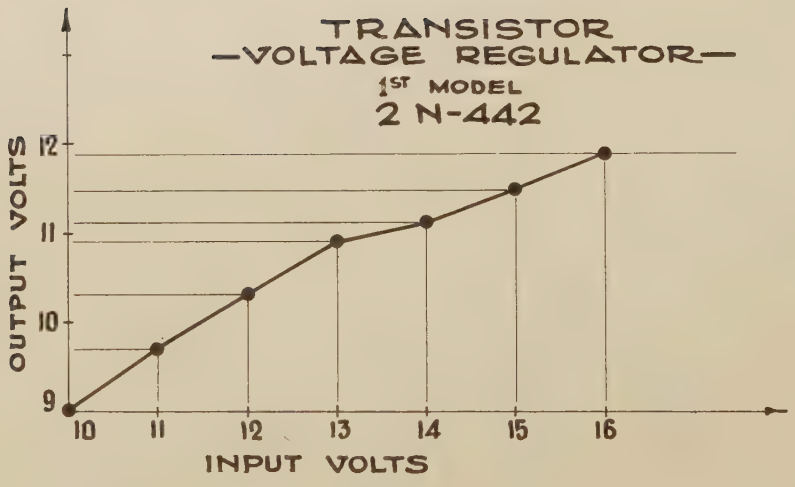
TRANSISTOR
—VOLTAGE REGULATOR—
1ST MODEL
2 N-442
OUTPUT VOLTS
12
11
10
9
10
11
12
13
14
15
16
INPUT VOLTS

Fig. 9.

A NEW MANUAL MOBILE TELEPHONE SYSTEM

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Summary

A new manual mobile telephone system has been designed around the requirements of operation by "land-line" telephone companies in the domestic public land mobile service. A specific objective has been a significant reduction in first cost and annual charges of the system to the operating companies. Achievement of this objective has resulted in a system which is expected to bring mobile telephone service to many areas which have previously been without it. This system makes mobile service feasible in many smaller exchanges which are typical of those served by the Independent Telephone Industry.

"Manual" Mobile Telephone Service

The term "manual" as applied to telephone service indicates the presence of a telephone operator to complete the call. Practically all mobile telephone service has until recently been offered on a manual basis. Various types of dial, or automatic, service are now becoming available in smaller communities.

The "manual" system described herein is designed to coordinate with and to be compatible with the existing nation-wide mobile telephone service offered by the Bell System and by many Independent Telephone Companies. Dial service in its present form is usually offered as a strictly local service in which only locally-registered mobile units may be contacted from the telephone exchange.

Service Requirements

Operating requirements of public telephone service differ from those of conventional two-way radio in a number of important respects. Among these is the need for nation-wide compatibility of the method of selectively signaling the mobile units. A uniform method of transmitting supervisory signals to the mobile service operator is also required. Another prime requisite is telephone-grade transmission, since a mobile service call may connect to any other telephone, anywhere. Speed of service and simplicity of operation are important, since the telephone subscriber is unwilling to make any concession to the fact that radio is the transmission medium.

At the telephone central office, the method of handling traffic must mesh smoothly with other tasks of the telephone company's traffic employees. The number of supervisory indications to be noted and understood by the switchboard operator must be held to a minimum. Methods of timing and ticketing calls must be comparable to those used on other classes of service.

From a technical operation standpoint, a high degree of standardization is desirable. This permits thorough and efficient training of maintenance personnel, who also handle many other types of equipment of comparable complexity. Trouble indications and routine operational functions must be presented in a manner which is consistent with established industry practice. Routine maintenance and trouble clearing must be clearly explained and easily accomplished in order to assure that the mobile system will be in a revenue-producing condition for a high percentage of the time.

Systems Engineering Approach

A mobile communications system designed to mesh smoothly with the fixed telephone network can be defined in terms of certain necessary elements. This breakdown is indicated in Figure 1. Necessary parts of such a system include a transmission facility, land-to-mobile connecting equipment, a mobile selective calling device, radio-to-telephone connecting and control equipment at the mobile unit, and r-f accessory equipment.

At the time this development was undertaken, there was no completely integrated system of this type available in the telephone field. There was, however, considerable equipment on the market which appeared to be applicable as components of such a system. The systems development approach followed was to evaluate each of these components in the light of its possible telephone application. This provided a starting point in the case of many of the system elements. It was then necessary to work with the various manufacturers to obtain detailed conformance to the requirements we

had in mind. This in most cases results in a special model of such equipment for use in the system. In some instances, engineering modifications have been worked out for application in our own shop to production equipment of other manufacturers, in order to assure exact conformance to the specifications established.

In the case of the land-to-mobile connecting equipment, no suitable supplier could be found and a development project was undertaken to provide this function. This equipment must be applied in a variety of different ways, depending upon the nature of the fixed plant and upon the coverage and traffic requirements of the mobile service area. Signaling and supervisory requirements are fixed by the practices already in widespread use. Therefore such equipment becomes a rather specialized item.

Land-to-Mobile Connecting Equipment

The supervisory, signaling and control functions built up over the years in "wired" telephone practice are not readily transplanted to the radio transmission medium. Great dependence has been placed upon the physical path and the ability to transmit low-information -content signals simultaneously with speech. This gives the telephone subscriber the service and operating features to which he is accustomed today. Providing this same service and operating features via radio is the task of the land-to-mobile connecting equipment.

A simplification of the problem is achieved in manual service by operating it along the lines of a local-battery exchange, as practiced in the telephone industry. That is, the subscriber must consciously initiate his request for the operator by pressing his push-to-talk button. (In local battery service, the subscriber makes use of the hand-cranked magneto for this purpose.) Dial mobile systems provide service more directly comparable to that available from automatic fixed telephones,[1] but such arrangements do not mesh with the existing nation-wide service.

Mobile telephone in all its forms is still party-line service, but fully-selective signaling is practiced. In manual service, this signaling follows a 600/1500-cps (cycles per second) sequential, frequency shift scheme. This signaling is standardized in the Bell System in various types of land and maritime services, including those operated in the h-f bands. Although somewhat slow considering present-day demands upon channel time, it has the advantage of a high degree of reliability. The various selective calling decoders developed for this type of signaling will usually decode the signal accurately, even on the circuits where commercial message service is no longer possible due to high noise.

The Type L-2 Control Terminal

The heart of the land-mobile connecting equipment is the base station control terminal. A block diagram of the Type L-2 Control Terminal is shown in Figure 2. This terminal, in its basic form, provides the functions:

1. Provides level-coordinating gain and loss elements in the audio line.

2. Provides test jacks in both the two-and four-wire lines, looking toward the switchboard and towards the base station radio equipment.

3. Provides a control and test position from which the technical operator can take over from the switchboard operator and operate or shut down the system.

4. Provides compositing arrangements in both directions to permit the d-c indications to be transmitted through the outside plant over the same pair with the voice.

5. Provides time delay circuitry to assure that the land transmitter has time to come on before signaling the operator to start dialing.

6. Receives interrupted d-c dial pulses from the switchboard and converts them to frequency-shift sequential signals for transmission over the mobile channel.

7. Includes sensing circuitry which converts an incoming Codan relay indication into a supervisory signal at the switchboard.

The control terminal was designed to use only transistors and telephone-type relays as dynamic elements. This permits operation from the standard -48 volt battery plant typically found in telephone offices. The dual-frequency oscillator and electronic switching circuits are mounted on printed circuit boards which plug into the basic panel. An extension cable permits easy testing of these units out of the panel.

Talker Level Considerations

Design of a radio system for telephone service must anticipate the sizeable range of volumes which will be encountered out of the fixed plant. In addition to the normal variation of talker volume between individuals, any service which is interconnected to the

telephone network must consider the variation in levels due to the many ways in which a telephone connection can be built up on the land end of the mobile service trunk. These possibilities range from a subscriber loop terminating in the same switchboard to a 4,000-mile toll call consisting of eight tandem connections.

This situation is illustrated in Figure 3. For design purposes, talker volumes of local subscribers are assumed to range from -8 to -31 VU (Volume Units) at the mobile service switchboard. A 17-db toll circuit has been assumed as typical of the loss to be encountered in a built-up connection terminating locally in a mobile service call, with the distribution of talker volumes still 23 db wide, from -25 to -48 VU. The 17 db is somewhat less than the maximum normally considered possible on toll connections, but some allowance is made for the probability that toll calls terminating in a mobile unit will not be made over longer distances. On the other hand, the range of talker volumes over toll circuits will probably be somewhat less than the 23-db assumption. Talkers over long toll circuits tend to a higher average volume and a smaller standard deviation, so that the -48 VU estimate is probably a conservative one.

The total range of 40 db in talker volumes is likely to exceed the capabilities of the radio employed. The system is standardized for operation on the narrow band channels required for all new service installed after November 1, 1958. This effects a certain reduction in the dynamic range possible and makes control talker volumes more important than in the "wide-band" applications.

Where the full range of talker volumes is expected from the land-line plant, use of the "AGAMP" (Automatic Gain Adjusting Amplifier) is considered desirable. The AGAMP maintains an output volume of 0± 4 VU with input variations of -8 to -48 VU. This permits the base transmitter to work more closely to its maximum peak deviation on all calls, with a corresponding increase in coverage.

Some telephone companies may anticipate that the majority of the land-line traffic will originate in the local exchange area, with very little coming from toll network. In such situations, a cost saving is possible by omitting the AGAMP. It is considered that the radio will be able to handle satisfactorily the 23-db range (from -8 to -31 VU) which has been assumed as coming from the exchange plant. Even with narrow band operation, present-day radio is considerably improved over that which was available at the time when it was considered necessary to employ an automatic gain adjusting device on all radio circuits. Also, toll circuits are being operated at lower net losses than they formerly were, and the figure corresponding to the 17 db assumed for this application was undoubtedly higher a few years ago.

Transmission from the mobile unit is not subject to the wide fluctuations which apply to transmission from the fixed plant. The loss between the handset and the mobile radio transmitter is always a fixed value. Even the talkers tend to be standardized in their volume output due to the nature of mobile subscribers and the conditions surrounding conversation from a vehicle. Therefore no special steps are taken to control talker volumes at the mobile unit.

Transmission Levels Through the Control Terminal

A level diagram of the base station control terminal is shown in Figure 4. In the example shown, a loss of 9 db occurs in the 2-wire line between the mobile service operator and the control terminal. The control terminal is shown at the same physical location as the base station radio equipment. Another possible arrangement contemplates extension on a four-wire basis between the control terminal and the radio.

The control terminal is arranged to re-broadcast all intelligence coming in from the receiver. This permits all users of the channel to monitor both sides of the conversation, in a true "party line" service. It also permits mobile-to-mobile, or "reverting" calls. This requirement introduces a level coordination problem which is solved by the "Reverting Call Level" Pad. It is necessary to maintain as high a level as possible out of the base station receiver to overcome possible line loss back to the central office. It is desirable to offer a low-loss path to speech coming towards the base transmitter from the land-line side of the connection for the same reason. This would normally result in an intolerably high level of re-broadcast speech at the transmitter input.

The Reverting Call Level pad is under control of the receiver codan relay, which in turn is under control of the mobile unit's push-to-talk button. It therefore comes into the circuit only while the mobile unit is transmitting. Thus a high loss is introduced into the path of the re-broadcast transmission, while the direct land-to-mobile transmission goes through without loss.

Electronic Switching Assembly

Standards for nation-wide mobile telephone service call for certain sequential events in the selective signaling of mobile units. One of these is to send one tone transition (600 to 1500 cycles) before starting to dial a number. This is equivalent to dialing a "1" and is done to clear all selective calling decoders which may be resting on advanced contacts due to noise or to previous signals. The decoders normally register a digit for each tone transition; therefore the dual-frequency oscillator must shift first in one direction and then the other with each pulse received from the

operator's dial. Finally, when dialing is completed, transmission of the last tone must be sustained for approximately four seconds to sound the audible signal in the subscriber's vehicle.

These functions are accomplished in the electronic switching assembly and time delay circuits of the 90101A control unit.

Upon receipt of the "transmitter on" supervisory signal at the control unit, relays operate to turn on the transmitter and to apply the oscillator output to the audio line. The same relays key the bi-stable switch to the 600-cps mode, and start the two-second time delay circuit. A simplified version of the time delay circuit is shown in Figure 5.

The quiescent conditions are such that the timing capacitor C is shorted, the transistor Q-1 is biased "on" by R-2 and R-3 and the diode D is back biased. The voltage at the junction of R-2 and R-3 is approximately -24 volts.

The timing cycle is started when the input contacts open allowing capacitor C to charge. The initial voltage at the junction of resistor R-1, capacitor C and the diode D is -48 volts. It approaches zero potential according to the relation

$$E = -48\epsilon^{-\frac{t}{RC}} \quad (1)$$

However, as the junction potential decreases it reaches a point at which the diode becomes forward biased (approximately -24 volts). At this time, R-1 and R-3 form a divider that biases Q-1 to cut off. This, of course, de-energizes the output relay.

Keying of the signaling oscillator is accomplished by means of a bi-stable transistor switch. A simplified version of the circuit is shown in Figure 6. As shown, the bias voltage for each transistor is derived from the collector voltage of the opposing transistor. Therefore if one transistor, say Q-1 for example, is conducting its collector is at a low potential due to the drop across the collector load resistor R_{C1}. As a result, Q-2 is biased to cut off because the base voltage is less than the emitter voltage. With Q-2 cut off, there is no drop across R_{C2} and Q-1 is biased "on." This is in agreement with the original premise and is a stable condition.

Should a positive "off" pulse be applied to C-1 when Q-1 is off or to C-2 when Q-2 is off, no switching results. However, the converse will result in switching the transistors to the opposite mode. Also, if C1 and C2 are tied together at the input side, there will be a transition for each input pulse. Each of these methods are used in the 90101A Electronic Switch.

The sequence of keying in the 90101A is as follows. When the "transmitter on" supervision is received, the transmitter is turned on, the two-second time delay is started and a pulse is applied to terminal 1 to insure that the initial signaling tone is 600 cps. After two seconds has elapsed, the two-second time delay circuit keys the electronic switch at terminal 2 to change the tone to 1500 cps. This in effect dials a "one" to reset all mobile units. The two-second time delay also provides the "start dial" indication to the operator and triggers the four-second time delay. Except for resistor-capacitor values the four-second delay circuit is similar to the two-second delay.

The operator has four seconds in which to initiate dialing. Each dial pulse keys the switching circuit via a separate input network which in effect ties terminals 1 and 2 together. Each dial pulse also resets the four-second time delay. Should four seconds elapse before dialing is started or a four second interval occur with no dial pulse, the time delay operates to key the switching circuit, thus sending out a final clearing tone. It also disconnects the signaling oscillator from the transmitter by means of a slow release relay. Thus, in the case of misdialing or abandoned calls, the control terminal automatically resets all mobile selectors and clears itself. In order to start up the signaling tone again, after losing it due to the four-second time delay, the operator dials a "1"which enables the dialing circuit for another four seconds.

Mobile Subscriber Station Equipment

In arriving at a standard assembly of mobile subscriber station equipment, emphasis was placed upon reliability, standardization and cost. Consistent with good transmission and otherwise satisfactory operation, it was considered desirable to choose radio equipment which is in widespread use and for which spare parts and up-to-date service information would be readily available. At this stage in the development of mobile telephone service, many telephone companies are contracting maintenance of their radio gear to outside organizations.

The telephone unit used in the subscriber station assembly is also typical of equipment in widespread use, for which maintenance spares are readily available in most telephone exchanges. This unit is shown in Figure 7. It includes a four-channel selector switch with a "channel-homing" feature to prevent lost incoming calls. Whenever the handset is replaced on the cradle, the channel homing relay returns the set to the channel of registry regardless of the position of the selector switch. In this manner the equipment is constantly on a standby position to receive calls even though the subscriber may inadvertently neglect to return the selector switch manually.

An electro-mechanical selective calling decoder

is standard with the system. This extremely rugged device is capable of a large number of combinations and can have its code changed readily. Various redundancy features reduce the possibility of false signals to an almost infinitesimal value.

Selective calling decoders widely used today normally register a digit upon the transition from one tone to another. Thus to register the digit "6" would require three transitions from 600 to 1500 cycles and three transitions from 1500 to 600 cycles, occurring alternately. A cost reduction can be achieved in the decoder by arranging it to register the pulses of only one of the two tones rather than the transitions. Extensive tests have indicated that this can be done without any loss in reliability or sensitivity. The penalty is taken in terms of a reduction in the number of combinations possible, since the decoder only steps once for each pair of transitions. Such a limitation will not be a serious one in most of the smaller exchanges where this system is expected to find considerable application. Where more numbering combinations are required (as in the case of "roving" subscribers) the two-tone decoder can be used on a directly interchangeable basis.

"Fixed Mobile" Subscriber Service

A "Fixed Mobile" subscriber station assembly has been engineered which permits fixed subscriber to share the radio party line with the mobile units on the same channel. Such operation is permitted in the rural radio service category on a secondary shared basis with mobile subscribers.

The fixed mobile equipment is similar to that used in vehicles except that it operates from 117-volts a-c, terminates in a conventional telephone instrument and uses a fixed, directional antenna. The telephone set is modified to provide for a key switch and for call lamp indications. A photograph of the manual fixed mobile telephone is shown in Figure 8.

Other Accessory Equipment

To provide a complete system has required considerable investigation into transmission lines, antennas, fittings, housings and other components. In most of these cases, commercially available items have proven to meet system requirements and have been grouped under appropriate nomenclature for convenience in engineering an installation into the telephone plant.

In this category, one deficiency which developed was the lack of an adequate large-size weatherproof housing for base station control and radio equipment. Outdoor mounting of control terminals, base station transmitters and receivers, and of satellite receivers will often be necessary in the telephone companies' applications.

Because of the requirement for rear access, the housings which have been used to mount carrier equipment on a 19-inch relay rack are not satisfactory. Accordingly, the Type 90110A Weatherproof Cabinet was developed. A photograph is shown in Figure 9. This housing uses "clamshell" doors to provide the maximum possible access to both sides of the equipment rack. The rear door is fully interlocked to prevent accidental contact with hazardous voltages. A mechanical interlock prevents opening the rear door without opening the front door first, so that the equipment will not be inadvertently taken out of service when only front-panel adjustments are required. A pair of thermostatically-controlled blowers draw air in through a filter at the bottom of the cabinet and exhaust it through the hood at the top. A total of 36 1-3/4-inch mounting spaces are provided on the 19-inch rack. Maximim projection to the front of the rack is 7-1/2 inches, and 8-1/2 inches to the rear. Power, telephone and r-f wiring enter through knockout ports at the bottom.

Future Prospects

The manual, compatible system described herein is one of a family of systems now being developed which is expected to broaden considerably the use of radio by telephone companies in providing their exchange-type service. It is no longer possible to think of telephone service in terms of wire lines. Microwave radio is playing an increasingly predominant part in the toll circuit mileage of the telephone industry. Maritime and certain other special services have long been offered by means of radio.

The mobile subscriber and the inaccessible, rural subscriber, are going to receive more and more attention from the telephone industry over the next few years. The economy of this country is such that a premium is placed upon mobility of the individual, and the immediate future is certain to see great emphasis upon mobility of communications to go with him.

A dial mobile telephone system has already been developed as a companion system to the one described herein. The dial system meets the needs of smaller communities for economical service where compatibility with the nation-wide signaling scheme is not a requirement. A "spectrum-stretcher" to help relieve crowded mobile channel by means of trunking principles is now in the planning stage. Eventually a compatible dial arrangement with nation-wide subscriber access will have to be agreed upon by the industry. The day is not too distant when mobile service will become as universal and economical as conventional "wired" telephone service is now.

References

1. R. P. Dimmer, "Automatic Mobile Telephone System," Communication and Electronics, No. 39 (November 1958) pp. 723-725.

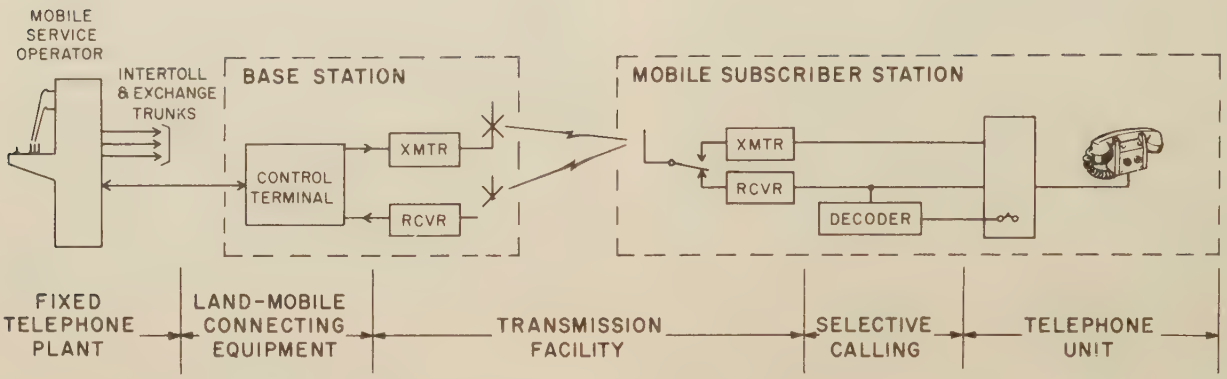

Fig. 1. Elements of a mobile telephone system.

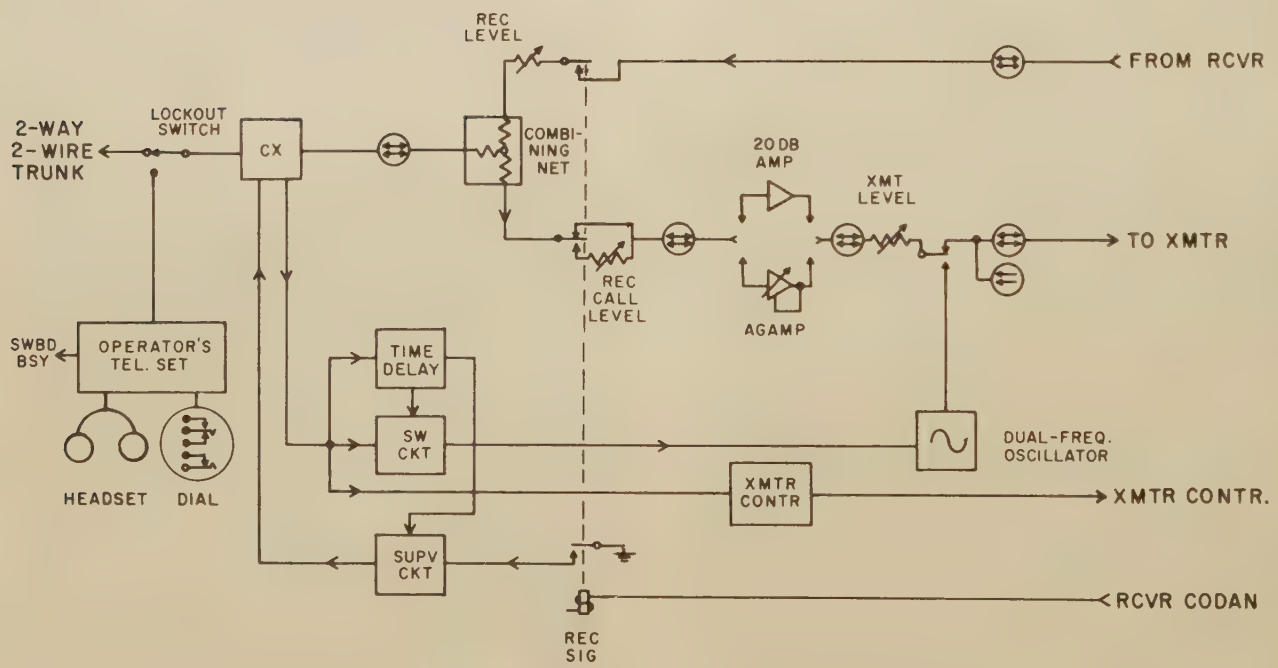

Fig. 2. Block diagram of the Type L-2 Control Terminal.

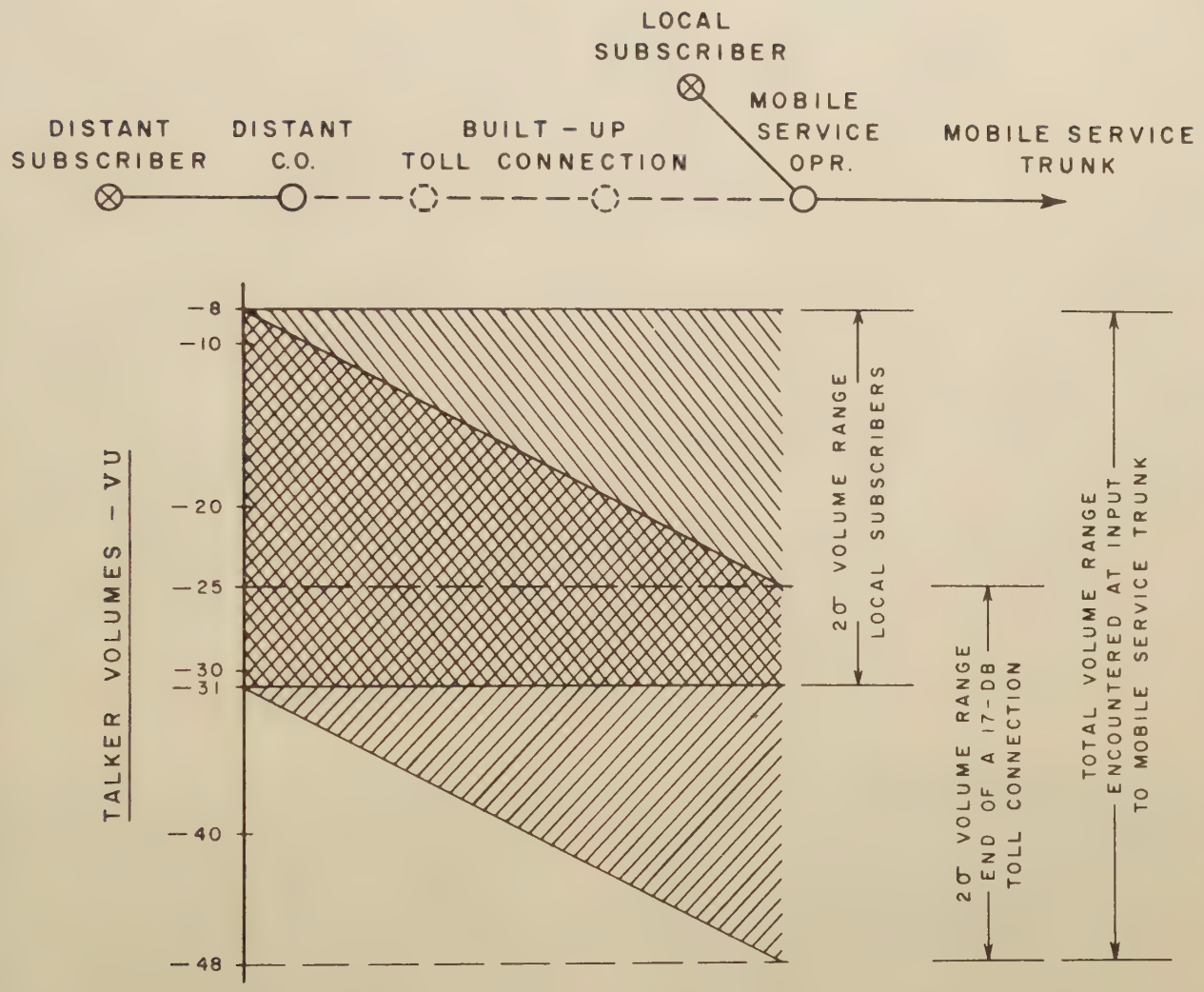

Fig. 3. Assumed range of talker volumes encountered in mobile telephone systems.

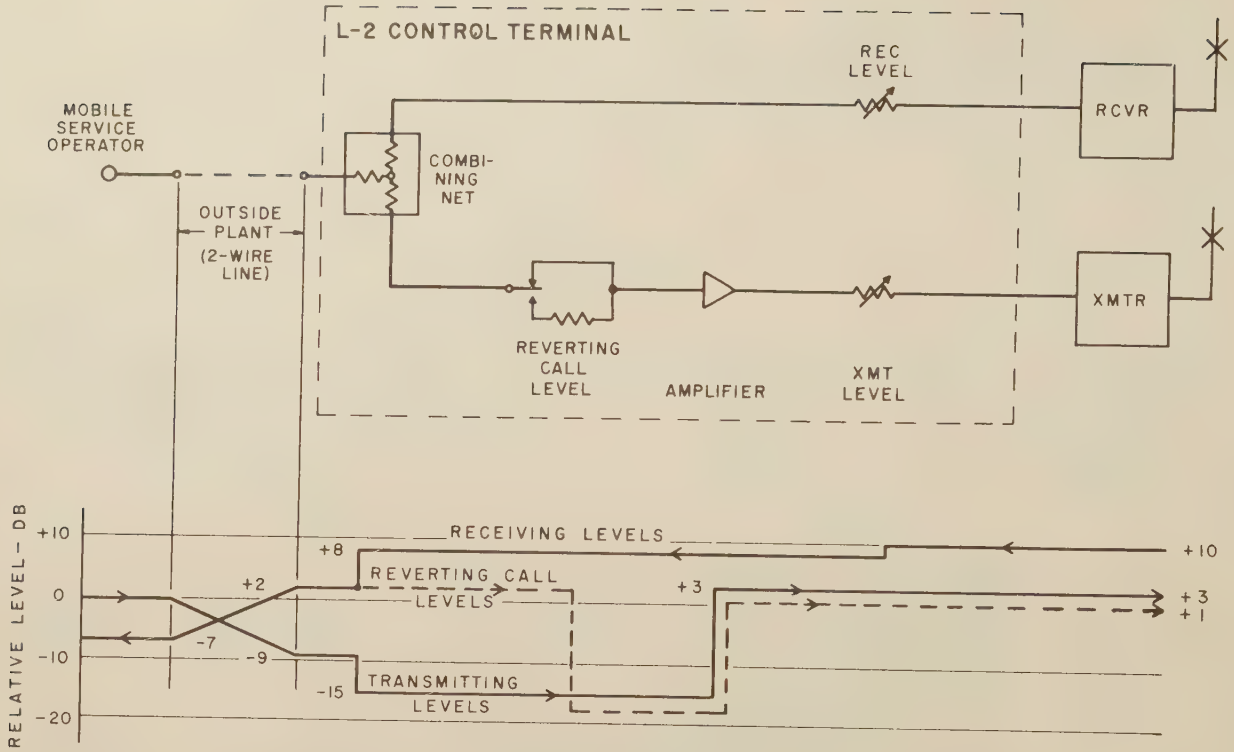

Fig. 4. Level diagram of manual mobile telephone system using Type L-2 Control Terminal.

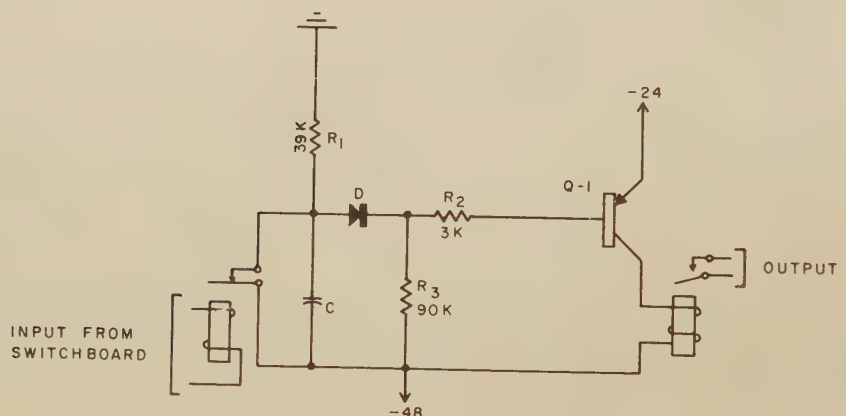

Fig. 5. Simplified schematic of 2 second time delay circuit.

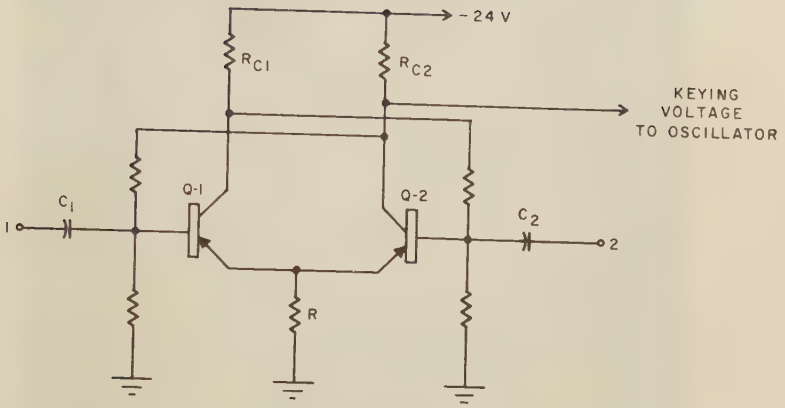

Fig. 6. Simplified schematic of bistable switch.

Fig. 7. Telephone unit for mobile installation.

Fig. 8. Telephone unit for "fixed mobile" installation.

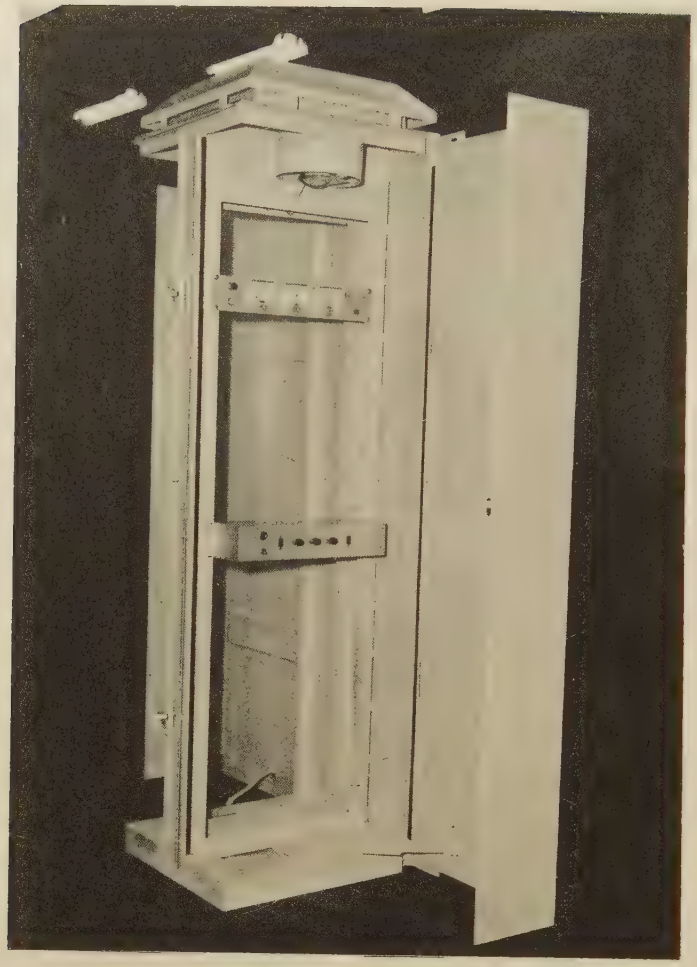

Fig. 9. Weatherproof housing for mobile equipment.

PERFORMANCE OF "LOW-PLATE-POTENTIAL" TUBE TYPES AT MOBILE-COMMUNICATIONS FREQUENCIES

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Abstract

The demonstrated advantages for hybrid-type automobile broadcast receivers of "low-plate-potential" tube types (those designed to operate at plate potentials of approximately 12 volts) has aroused considerable interest in the usefulness of such tube types at mobile-communications frequencies. This paper analyzes the performance of low-plate-potential tube types at frequencies above 2 mc. It also discusses performance requirements and design considerations for mobile-communications receivers and presents operating data on an experimental hybrid-type mobile-communications receiver using available low-plate-potential tube types. These data are compared with typical receiver-performance specifications for mobile service.

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